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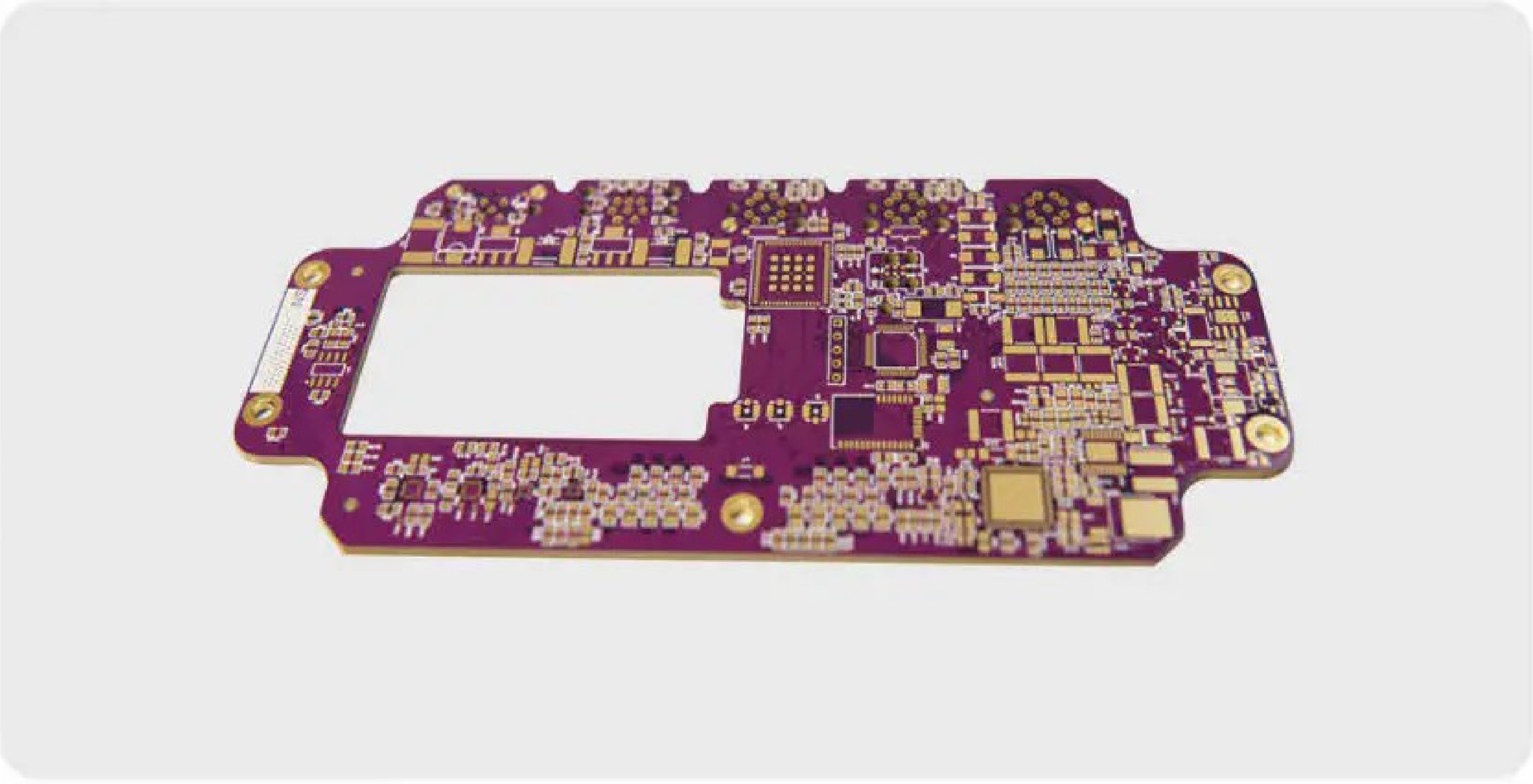

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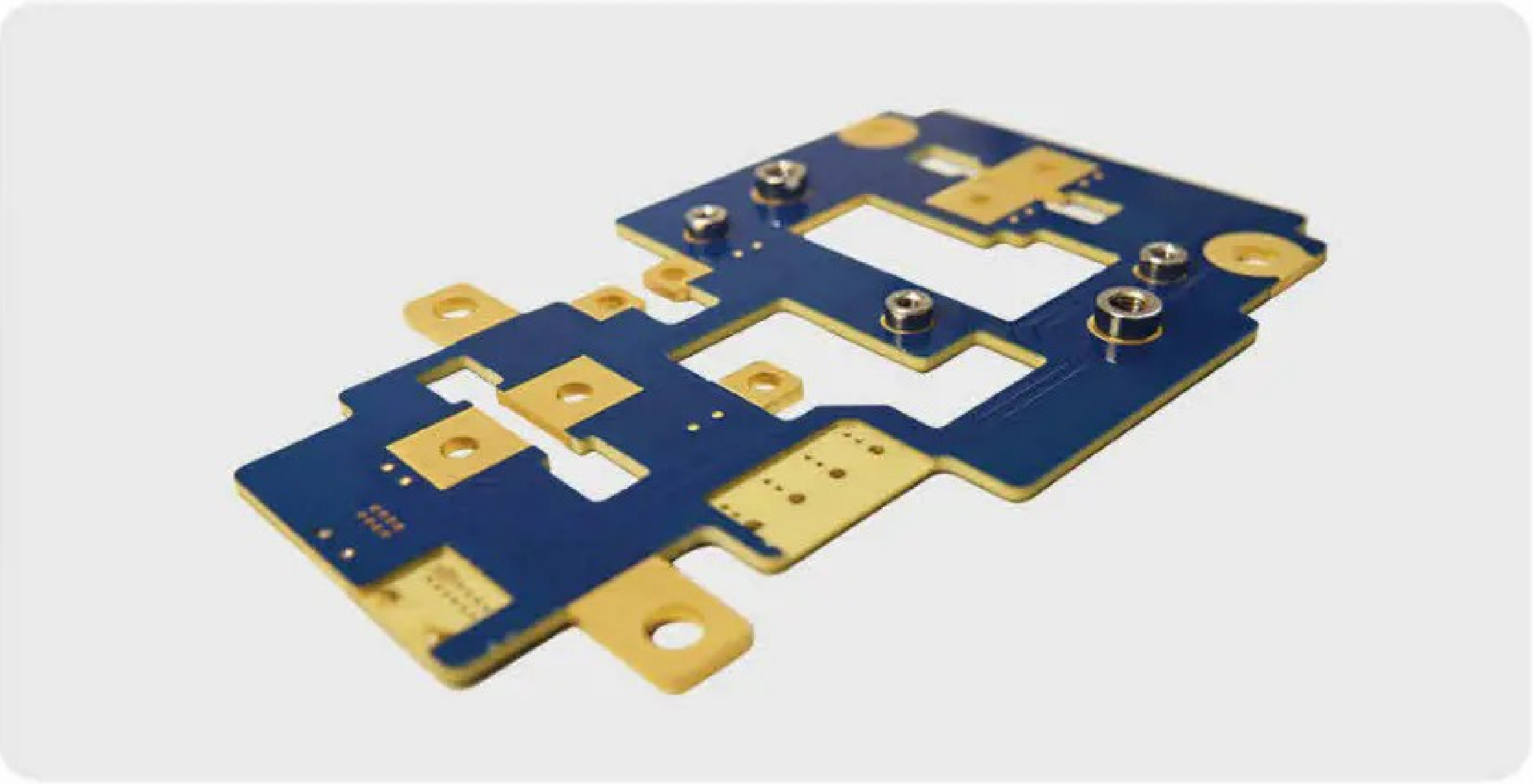

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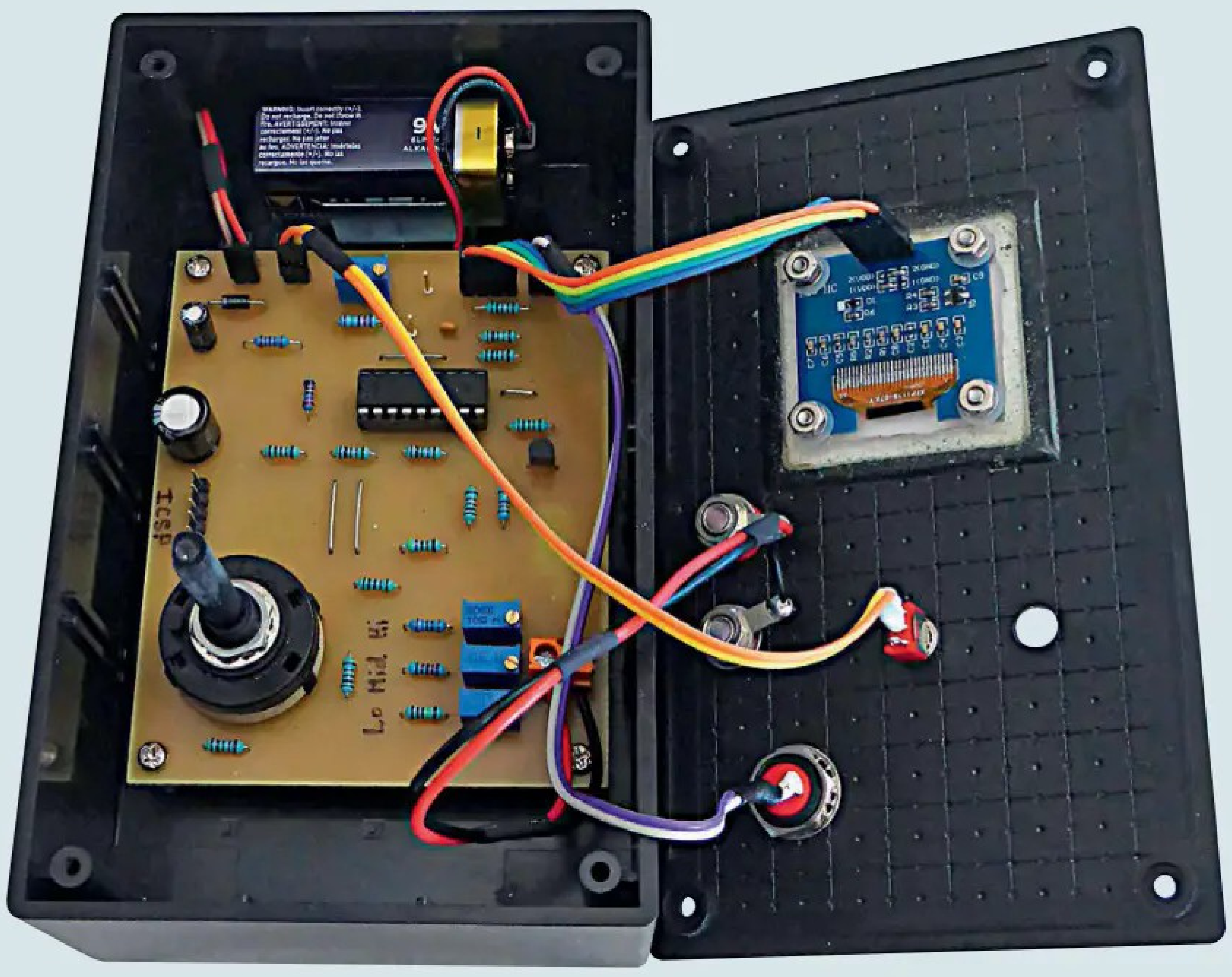

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Editorial offices
Electron Publishing (Australia)
Tel +61 2 9939 3295
Email pe@pemag.au
Web www.electronpublishing.com

Address mail to:
Electron Publishing (Australia)
PO Box 194, Matraville NSW 2036
Australia

Advertising enquiries
+61 2 9939 3295 pe@pemag.au

Editor Nicholas Vinen
Publisher Nicholas Vinen
Digital subscriptions Stewart Kearn **Tel** **07918 614662**
Online Editor Alan Winstanley
Web Systems Kris Thain
Production Bao Smith
Technical staff Tim Blythman, John Clarke

Print subscriptions
Practical Electronics Subscriptions
PO Box 6337
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Tel 01202 087631
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The lost art of backward compatibility

Part of the reason that Linux is still around after nearly 35 years (it's gaining in popularity, too) is the very sensible philosophy of "don't break user space". This means that when people are adding features to the Linux kernel or making other changes, they shouldn't prevent pre-existing software from running on the computer.

This is also what made DOS and later Windows so popular; you could mostly keep using your software even when you upgraded to a newer version of DOS/Windows because they had very good backward compatibility. And yes, Apple does the same with its computers, even going so far as to use processor emulation to keep older code running on machines using totally different CPUs.

However, embedded systems like Arduino do a relatively poor job of this. It's quite common to upgrade a critical library like an LCD driver, real-time clock interface and so on, and find that your program no longer compiles. This is usually because they decided that their old API was bad in some way and they changed it without regard to breaking existing code.

I think this is a real barrier to the adoption of Arduino and other similar systems. Nobody wants to write and test code, deploy it, then find later on that it suddenly no longer works. It's especially annoying when we're trying to teach people how to use these systems with example code, as often, they will complain that our examples don't work. And they are quite right!

What we wrote and published was valid at the time, but now the goalposts have been moved, and the compiler throws up its arms at this suddenly invalid code.

I think there needs to be a real shift in attitude among software engineers with a recognition that this sort of thing is not acceptable. By all means, they should improve their software interfaces, but they need to find a way to do it that lets old code be. If you don't like the interface the old function had, create a new one. The legacy function can be a 'wrapper' that simply calls the new function with the new syntax.

It would be like if you bought a new car, and this year Toyota decided that you should steer with your feet and actuate the accelerator and brake with your hands. We sensibly keep car controls mostly the same from year-to-year, with new controls being added on in other places. Software should adhere to the same philosophy.

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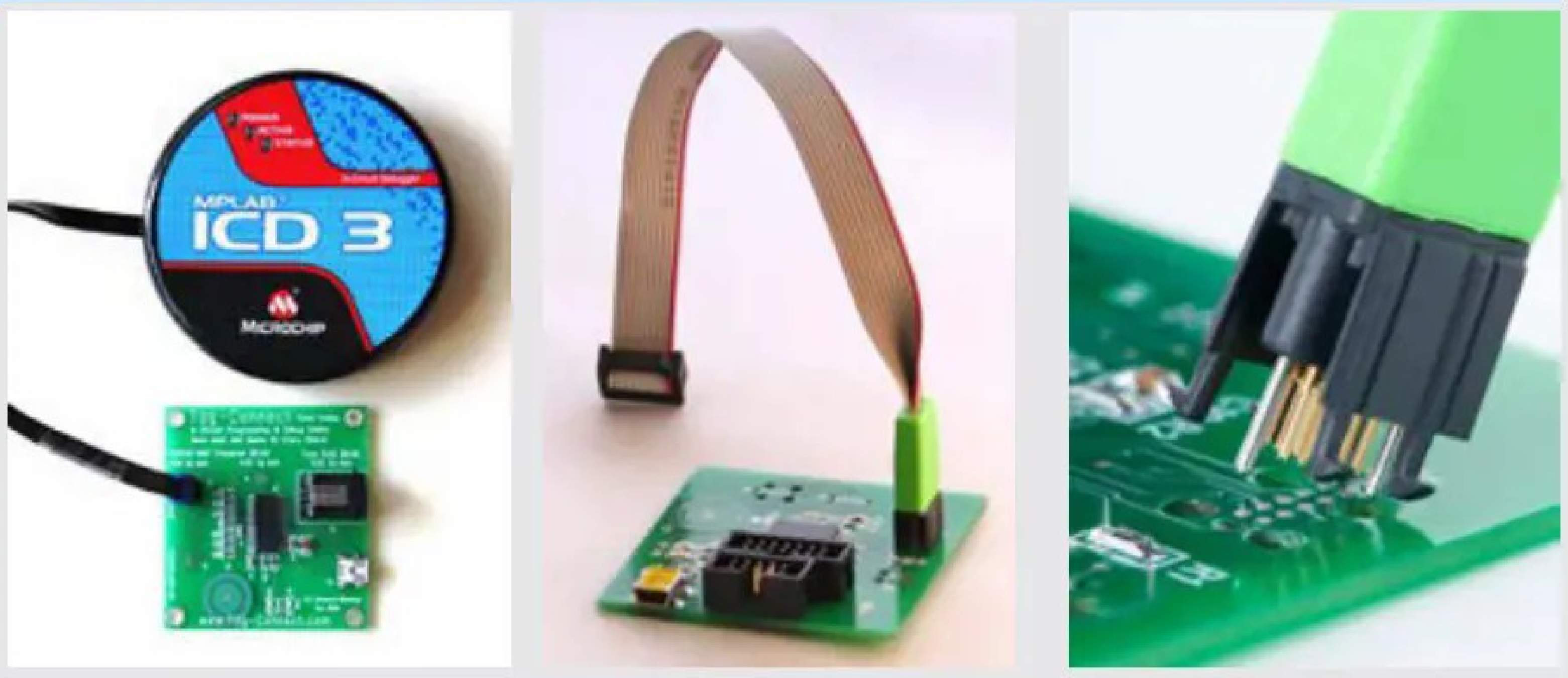

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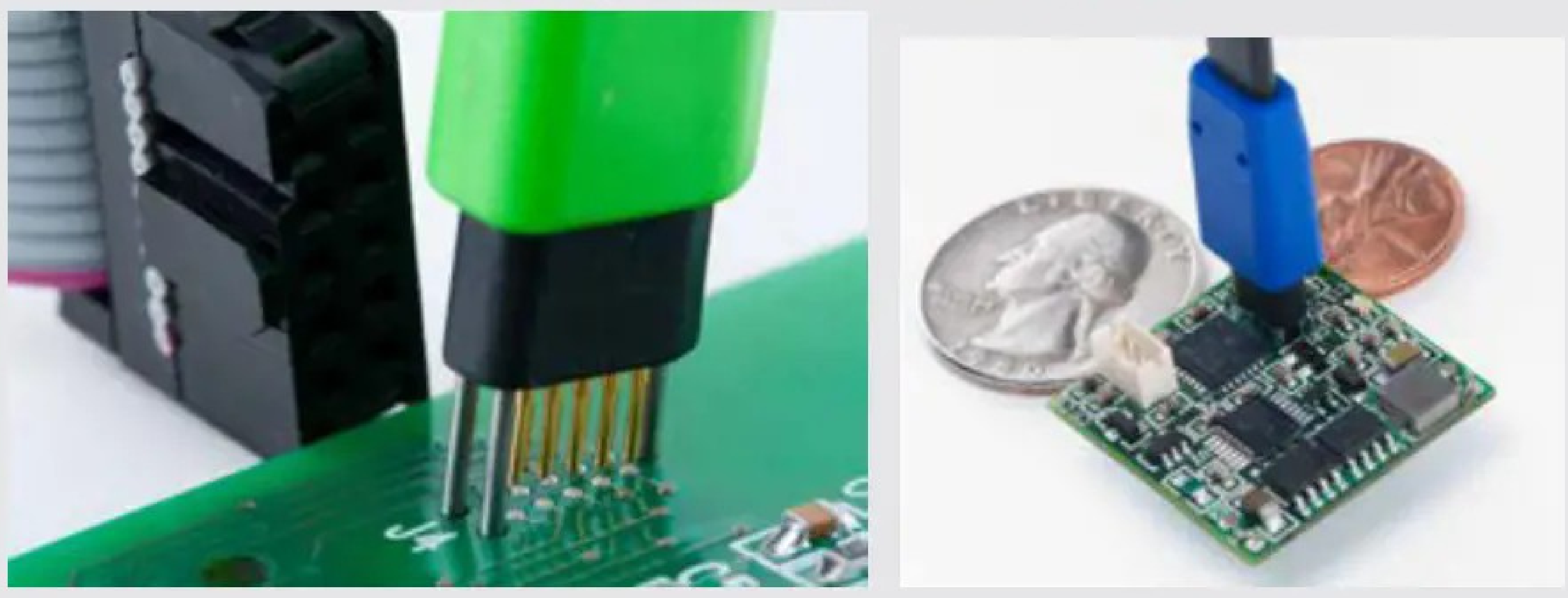

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Teach-In 2026

by Mike Tooley

World of Wireless – An Introduction to Radio and Wireless Technology

Part 2 – Introducing RF components and circuits

In the first instalment of this series, we began by setting the historical stage, highlighting major milestones and scientific breakthroughs in radio before moving on to explain the propagation of radio signals.

We provided an overview of the radio frequency (RF) spectrum, which spans from tens of kilohertz (kHz) up to hundreds of gigahertz (GHz), and examined how signals travel as ground waves at lower frequencies and sky waves at higher frequencies, with the aid of layers present within the ionosphere.

Our Hands-On project involved the construction of a very simple medium-wave AM receiver for headphone reception. This month, we start by examining a simple radio frequency circuit before moving on to explore the principal components on which it is based. Our practical project features an improved portable medium-wave AM receiver.

A VHF signal source

To begin, let's look at a simple RF circuit: a VHF signal source set to 70.4MHz, in the four-metre amateur band. The source delivers a stable, precise output a little below 10dBm (10mW) into a 50Ω load. It is ideal for antenna adjustment and receiver testing.

Instead of generating the final 70.4MHz signal directly, we use an inexpensive quartz crystal oscillator at one-tenth that frequency and then multiply it by 10 in a frequency-multiplier stage.

Fig.2.1 shows the circuit of the four-metre signal source circuit. It uses two 2N7000 N-channel Mosfet transistors (TR1 and TR2) as a fundamental-mode oscillator, and a third 2N7000 (TR3) as the ×10 frequency multiplier.

Frequency accuracy comes from a 7.04MHz quartz crystal, easily sourced online. The series trimmer capacitor (TC1) allows fine adjustment to account for component and temperature variations.

The tenth harmonic of the main oscillator is selected by means of a parallel-tuned circuit, L1 and C4, with the inductor made from five turns of tinned copper wire with the output coupling capacitor, C5, tapped at 1.5 turns from the 'cold' (i.e. supply) end. To permit tuning of the correct harmonic, L1 is fitted with an adjustable ferrite core.

The three 2N7000 transistors are low-cost enhancement-mode N-channel Mosfets (to be explained in more detail later). They are primarily designed for use in switching applications, but with appropriate bias arrangements, they can be used for linear as well as Class-C operation. They are suitable for output power levels of up to several hundred milliwatts.

The whole circuit is powered by a 9V PP3 battery and enclosed in a screened die-cast metal enclosure. The output signal is shown in Fig.2.2 using a software-defined radio (SDR) receiver.

Tuned circuits

Tuned (or resonant) circuits are used to discriminate between signals at different frequencies. Fig.2.3 shows two basic configurations for a tuned circuit: series and parallel. The frequency at which both are resonant, f_0, is given by the relationship $f_0 = 1 \div (2\pi\sqrt{LC})$.

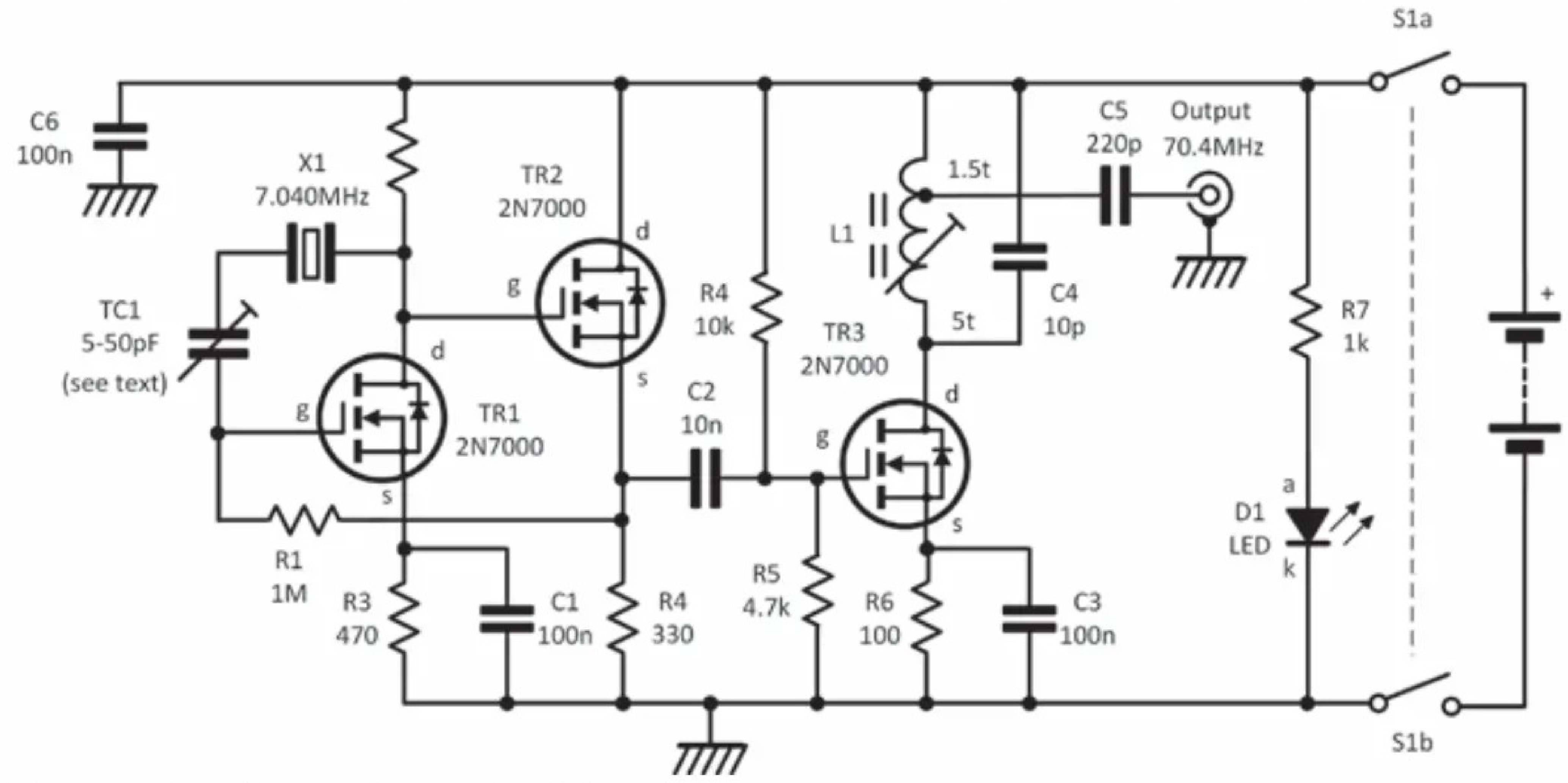
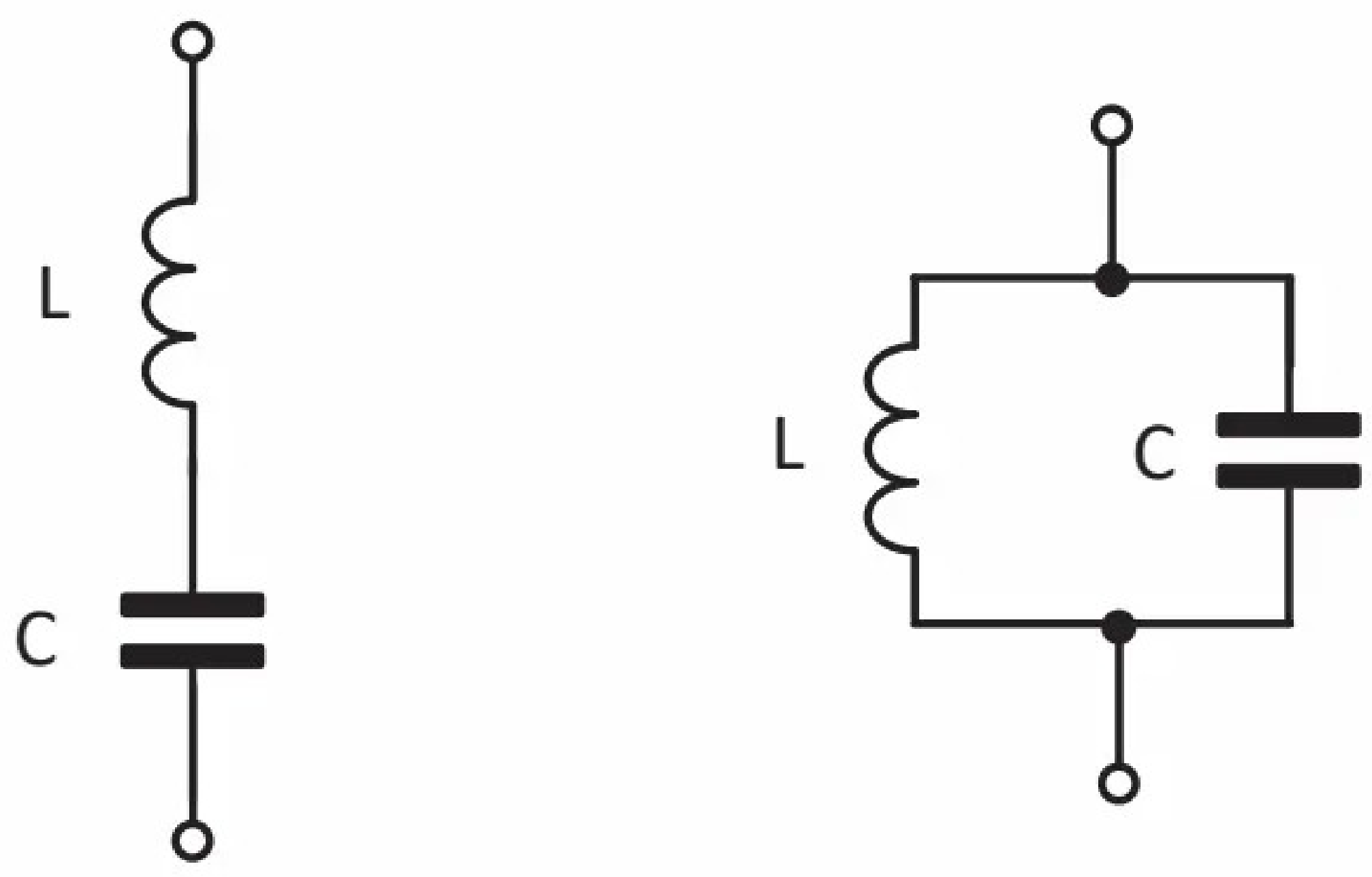


Fig.2.1: the circuit of our 70.4MHz signal source.



(a) Series

(b) Parallel

Fig.2.3: ideal series and parallel resonant circuits (an 'acceptor' and 'rejector').

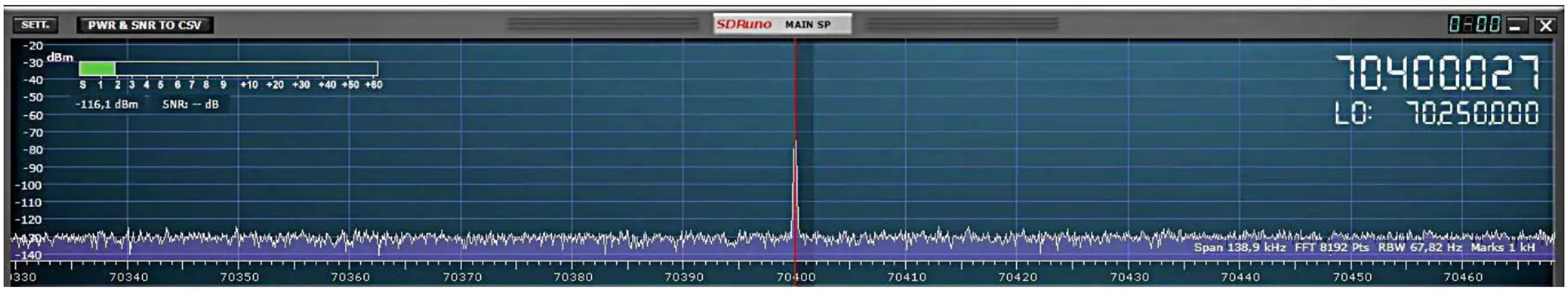

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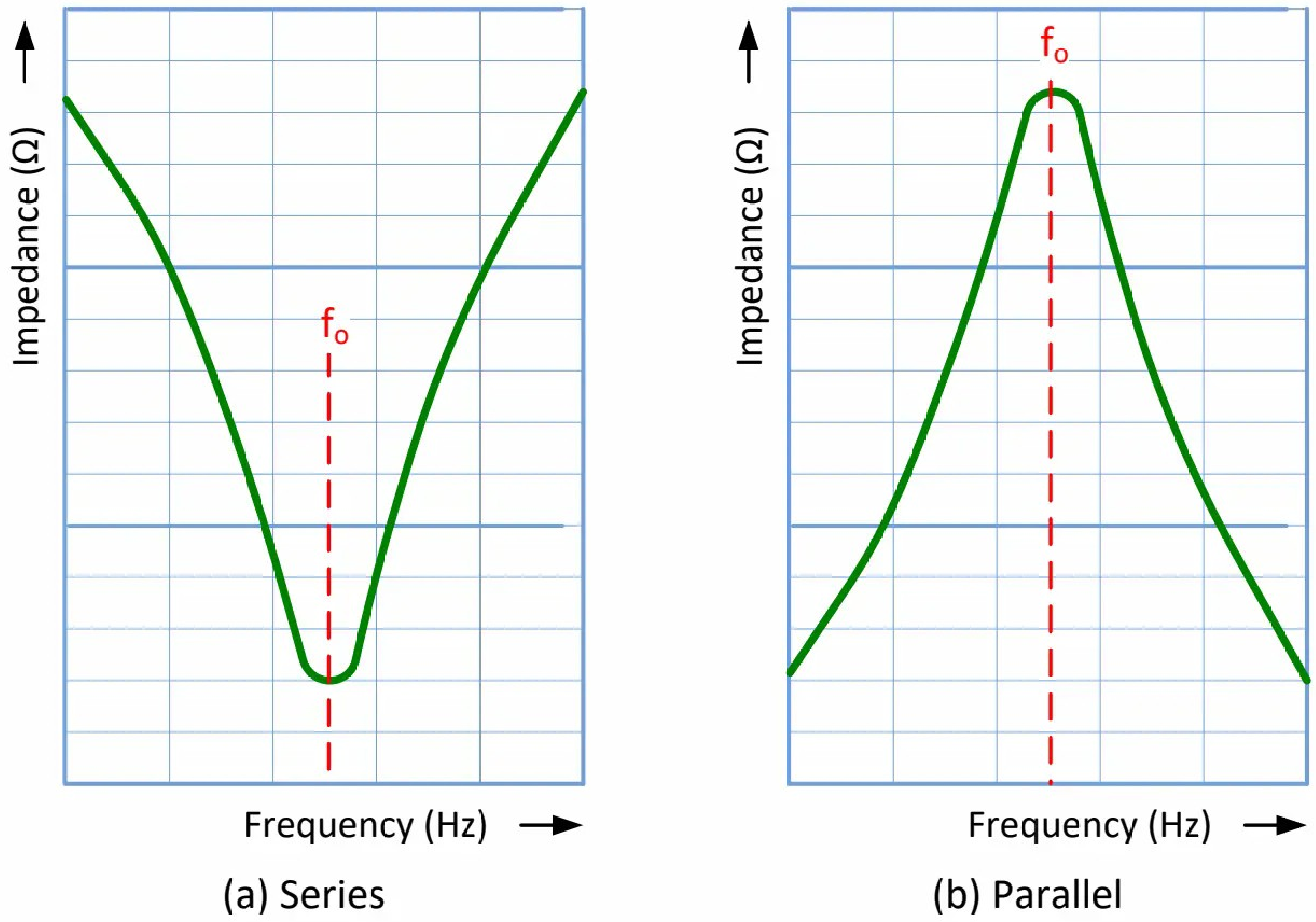
Fig.2.2 a panoramic waterfall (spectrum, ie, power vs frequency) display of the 70.4MHz signal.

All practical resonant circuits have losses, the most prevalent of which is that due to the series resistance associated with the winding of the inductor. Further losses occur at RF but, for simplicity, we will group these losses together and show them as a single series loss resistance, R_{Loss}, as shown in Fig.2.4.

The impedance-frequency characteristics of series and parallel resonant circuits are shown in Fig.2.5. It is important to note that the impedance of the series-tuned circuit falls to a very low value (ideally zero) at the resonant frequency; for a parallel-tuned circuit, it increases to a very high value (ideally infinite) at resonance.



(a) Series

(b) Parallel

Fig.2.5: the impedance-frequency characteristics of series- and parallel-tuned circuits.

For this reason, series-tuned circuits are sometimes known as acceptor circuits, while parallel-tuned circuits are sometimes called rejector circuits.

The frequency response (voltage plotted against frequency) of a typical parallel-tuned circuit is shown in Fig.2.6. This characteristic shows how the signal developed across the circuit reaches a maximum at the resonant frequency, f_0.

The range of frequencies accepted by the circuit is normally defined in relation to the half-power (-3dB power) points. These points correspond to 70.7% of the maximum voltage, and the frequency range between these points is referred to as the -3dB bandwidth of the tuned circuit.

Losses present in a tuned circuit (or imposed on it by the external circuitry to which it is connected) reduce its quality or Q-factor. Thus, as losses increase, the Q-factor is reduced, and vice versa.

The Q-factor also determines the selectivity of a resonant circuit. A higher Q-factor is associated with narrower bandwidth and thus increased selectivity, whereas a lower Q-factor is associated with wider bandwidth and less selectivity, ie, Bandwidth = $(f_0 \div Q)$ Hz.

For the series-tuned circuit shown in Fig.2.4, $Q = (2\pi f_0 L) \div R_{Loss}$.

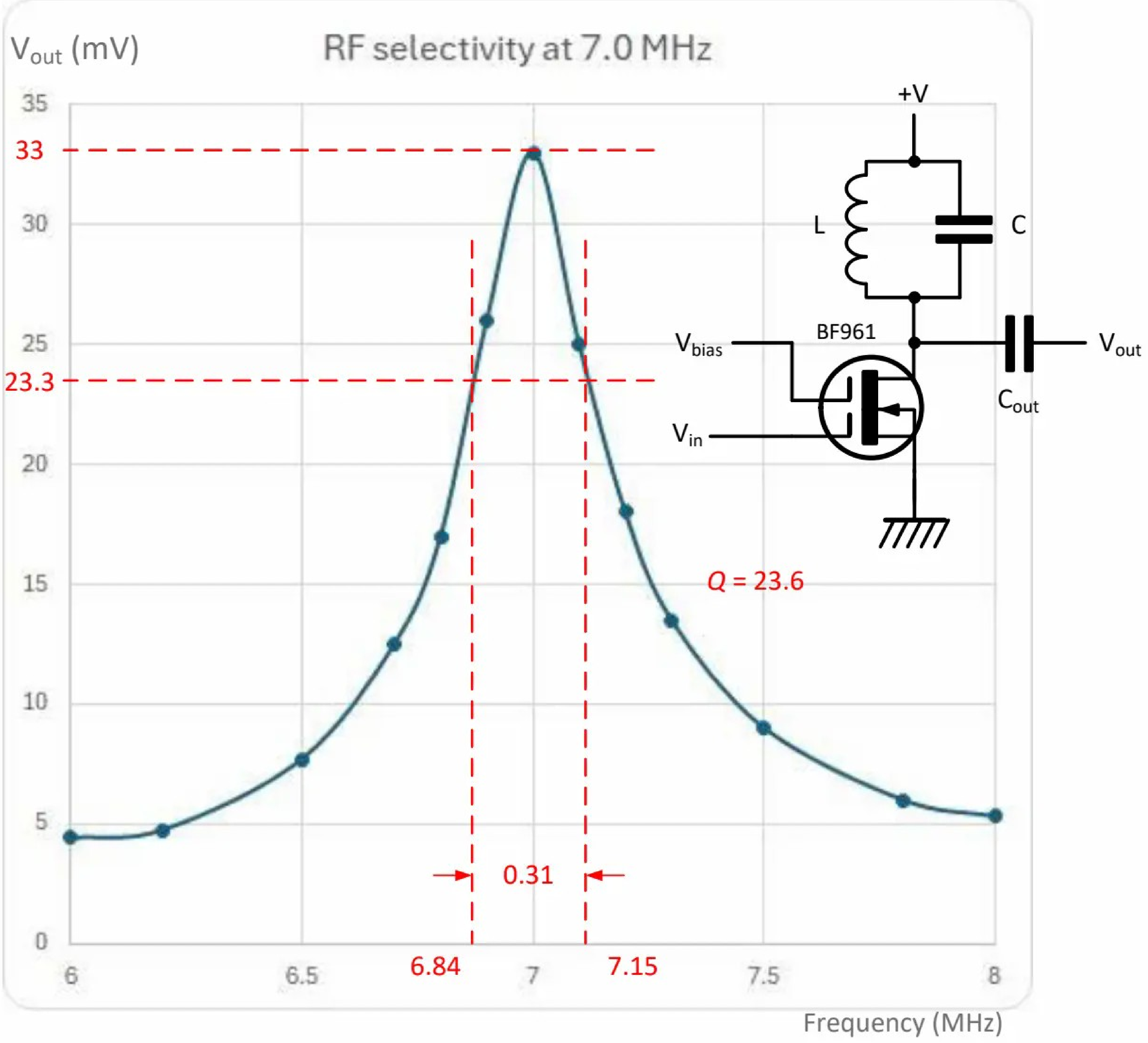
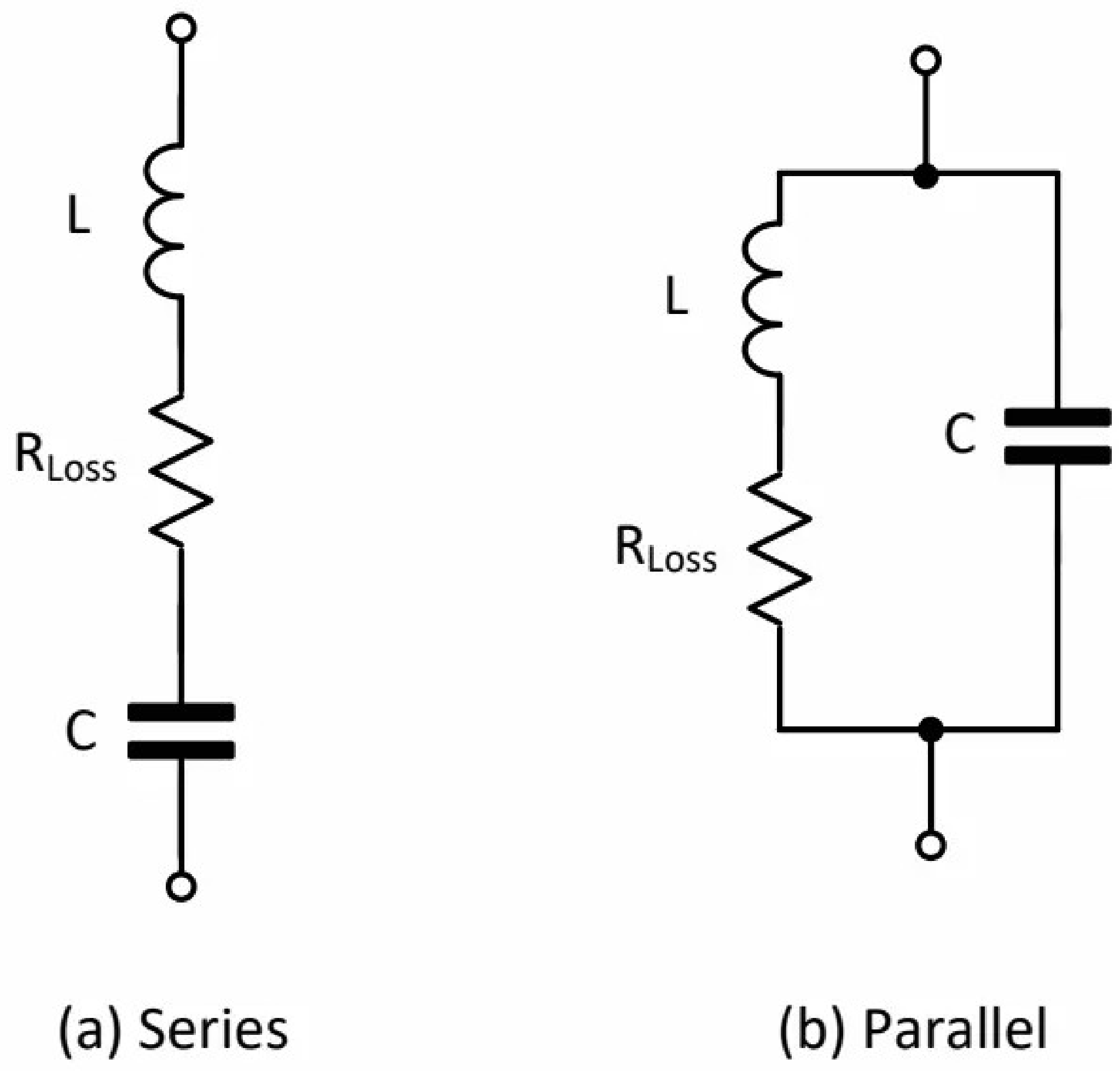


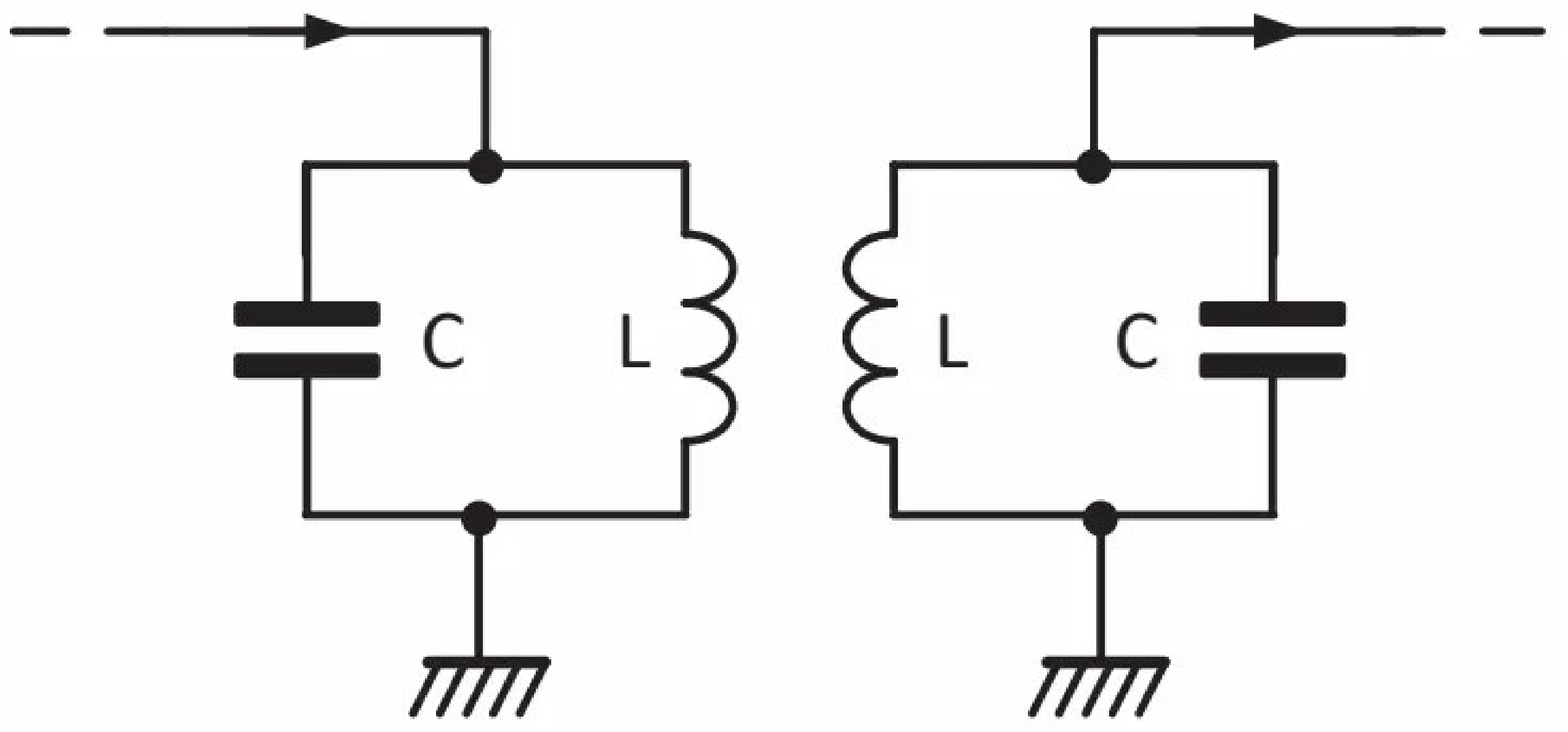
Fig.2.6: the frequency response of a parallel-tuned circuit in a 7MHz RF amplifier.


L
R_{Loss}
C
L
C
R_{Loss}

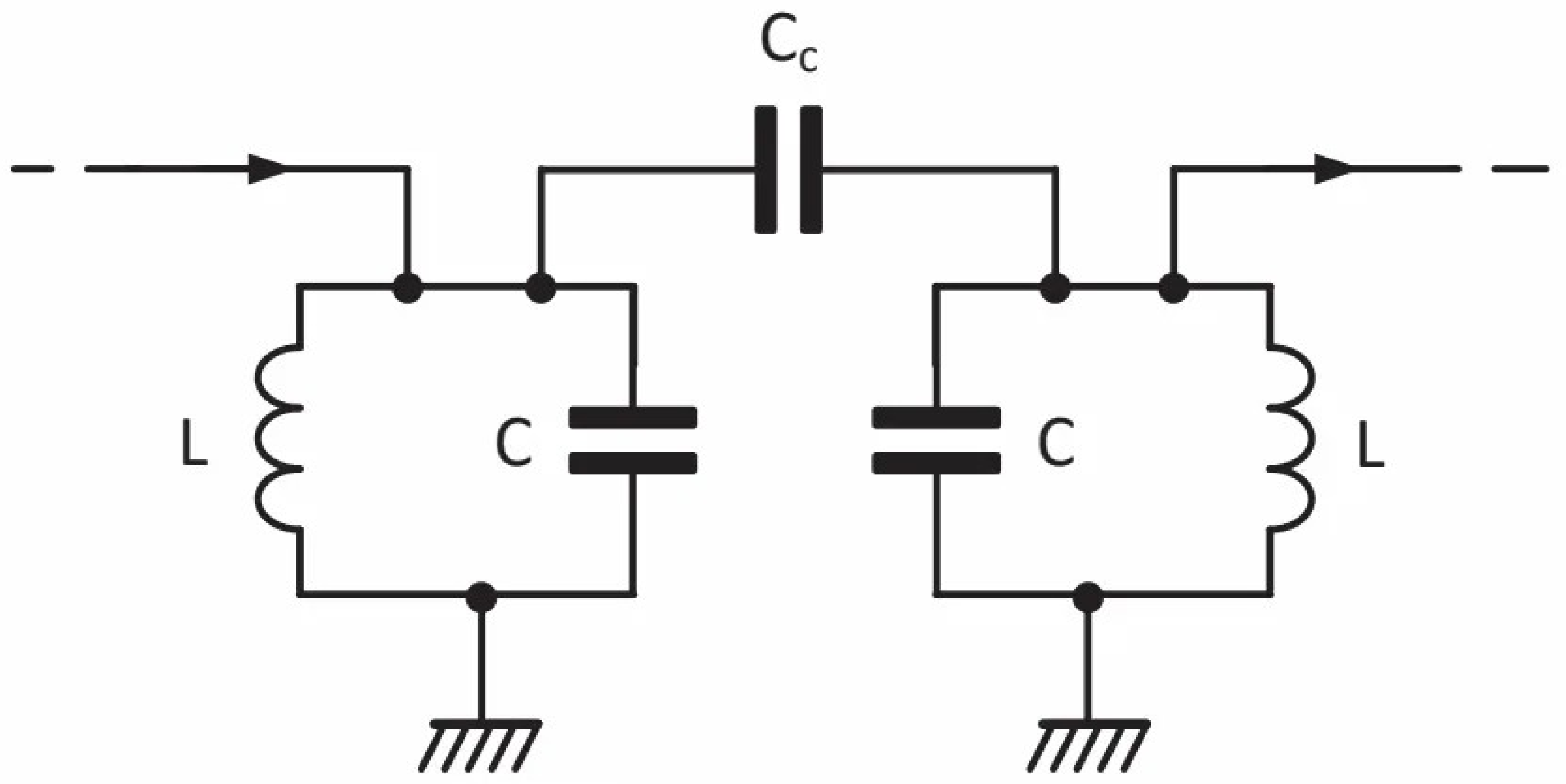
(a) Series

(b) Parallel

Fig.2.4: real tuned circuits have losses due to resistance and other factors.



(a) Inductively coupled



(b) Capacitively coupled

Fig.2.7: some coupled tuned circuits.

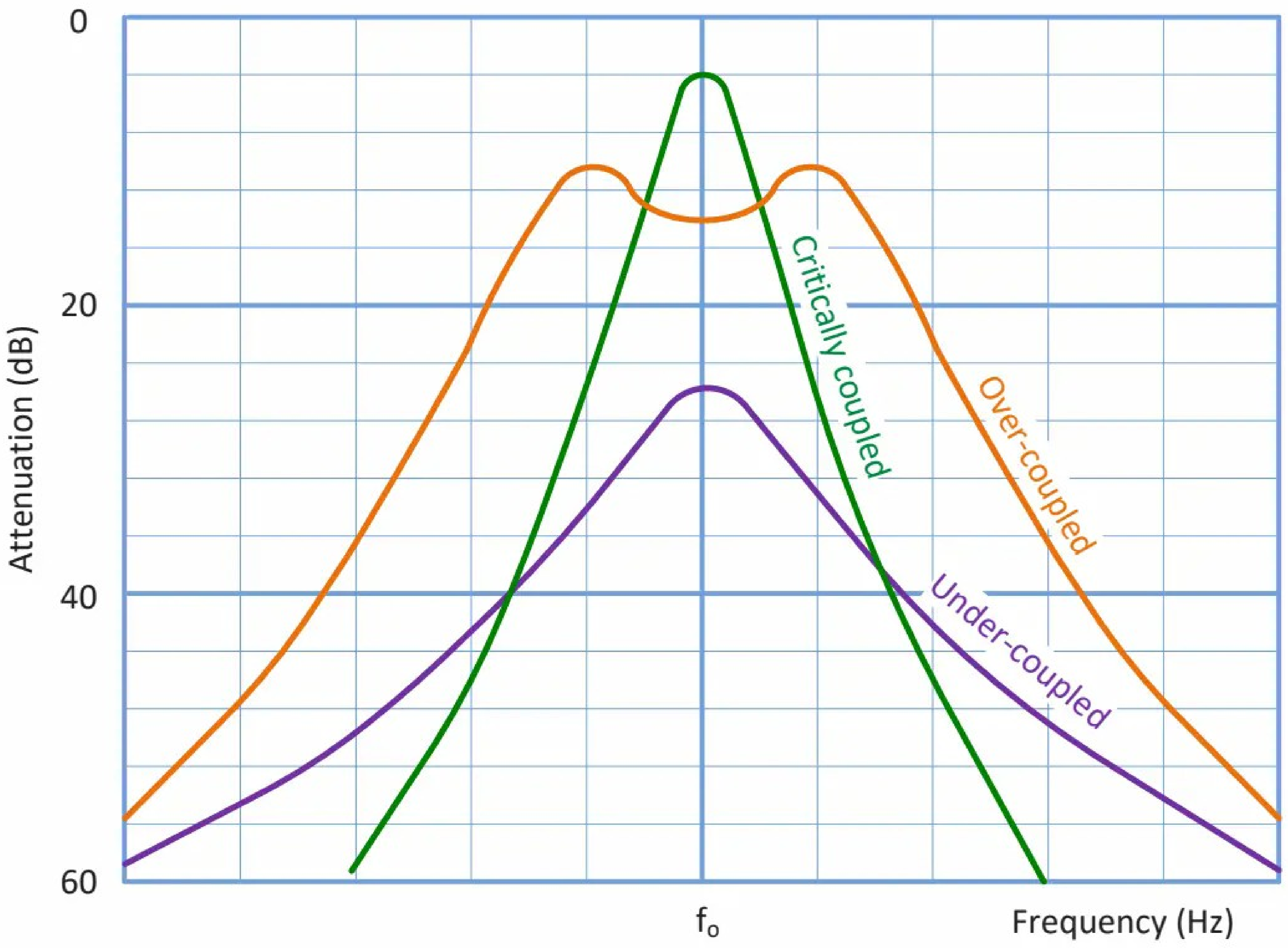


Fig.2.8: the frequency response of coupled tuned circuits.

RF filters and selectivity

A perennial problem with the design of receivers and transmitters is the need to separate wanted signals from other signals on adjacent frequencies. We refer to this as selectivity. A single RF tuned circuit will normally exhibit a Q-factor between 50 and 110. Selectivity can be improved by using multiple tuned circuits, but this brings with it the problem of maintaining accurate tuning of each circuit.

A band-pass filter can be constructed using two parallel-tuned circuits coupled inductively or capacitively, as shown in Fig.2.7. The frequency response of this type of filter depends on the degree of coupling between the two tuned circuits.

Optimum results are obtained with a critical value of coupling (see Fig.2.8). Too great a degree of coupling results in a double-humped response, while too little coupling results in a single peak in the response curve, accompanied by a significant loss in signal. Critical coupling produces a relatively flat passband characteristic, with a reasonably steep fall-off on either side of the passband.

Quartz crystals

Quartz crystals provide us with an alternative to conventional L-C circuits and have several applications in RF circuits. Not only can they be used to accurately determine the frequency at which a circuit oscillates, but they can also be arranged to form highly selective filters.

Quartz crystals usually consist of a thin slice of quartz (hexagonal crystalline SiO_2) with film electrodes of gold or silver deposited onto opposite sides. A fine supporting and connecting wire is soldered at a nodal point on each electrode, and the complete assembly is enclosed in an evacuated glass or metal envelope. Lead-out pins or wires facilitate connection with external circuitry.

The type of encapsulation, size, dimensions and pin spacing varies from one type of crystal to another. Fig.2.10 shows a selection of commonly available quartz crystals supplied in different encapsulations.

The electrical equivalent circuit of a quartz crystal is shown in Fig.2.11. The device can be considered either series or parallel-tuned depending upon which of the two capacitors, C_s or C_p, is allowed to become dominant by external circuit conditions.

The impedance/frequency characteristic of a typical quartz crystal is shown in Fig.2.12. Both series and parallel resonant peaks are evident; however, the frequency separation of the two is extremely small.

At the series resonant point, the crystal behaves as a pure resistance, and any change in external capacitance will have little effect. At the parallel resonant point, the impedance of the crystal becomes inductive, and a change in the external circuit reactance will have the effect of 'pulling' the crystal frequency away from its natural parallel resonant frequency.

Hence, if a crystal is to be used in a circuit that requires it to operate at parallel resonance, the load capacitance must be accurately specified. It is also worth remembering that the load capacitance is the dynamic capacitance of the total

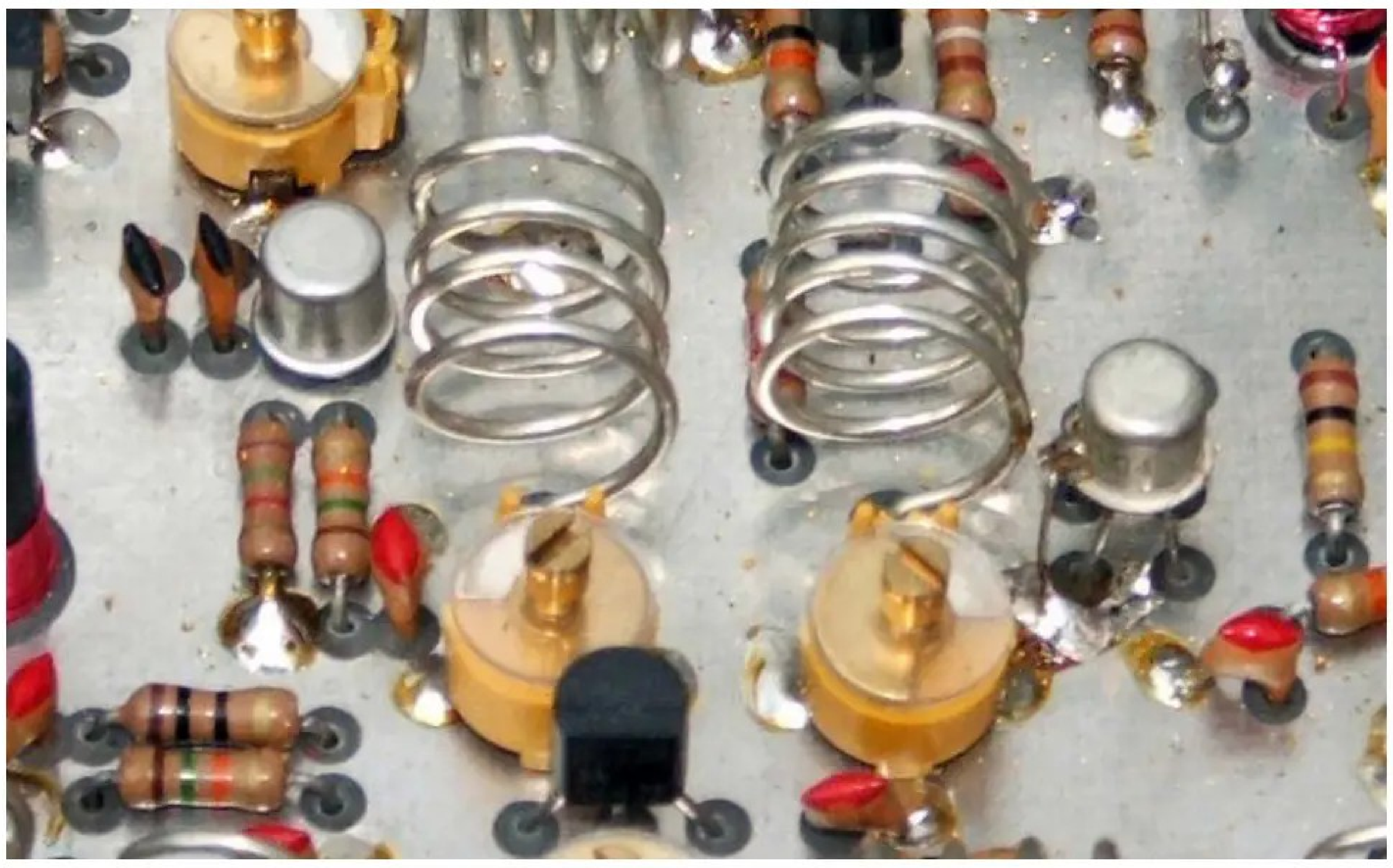


Fig.2.9: inductively-coupled tuned circuits in a VHF amplifier. The two trimmers are adjusted to get an optimal band-pass characteristic.

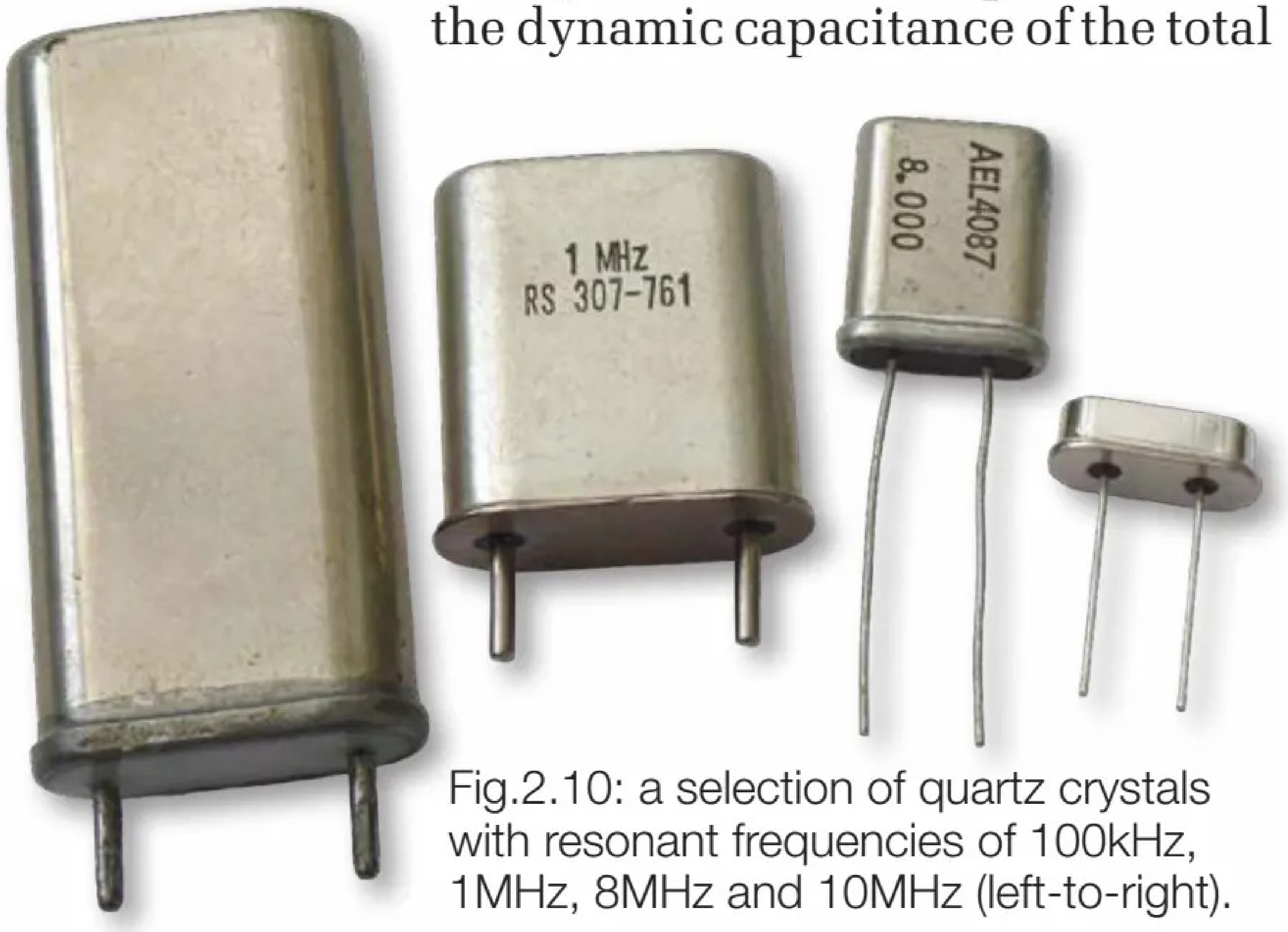


Fig.2.10: a selection of quartz crystals with resonant frequencies of 100kHz, 1MHz, 8MHz and 10MHz (left-to-right).

circuit measured across the crystal terminals.

In parallel circuit design, the load capacitance should be selected to operate the crystal at a stable point on its frequency characteristic (usually as close to the series resonant frequency as possible).

Quartz crystals exhibit extremely high Q-factors, which are many times larger than those that can be obtained with even the very best L-C tuned circuits. The reason for this is that the ratio of equivalent inductance, L, to series loss resistance, R, is exceptionally high.

Crystals manufactured for fundamental operation are designed to oscillate at their basic resonant frequency, whereas those intended for overtone operation oscillate at, or very near, an integral multiple of their fundamental resonant frequency. Generally, the third overtone is preferred, although fifth, seventh, and even ninth overtone devices are available.

At high frequencies, crystals become extremely thin, and are consequently more difficult and more expensive to manufacture. Thus, fundamental crystals are normally used at frequencies up to about 20MHz. Beyond this, overtone units are usually specified.

The circuit load condition must be known to ensure the correct mode of operation (series or parallel). In series mode, the quartz resonator is operated in a low-impedance state. In the latter, it is operated in a high-impedance parallel-mode state.

For parallel resonant crystals operating at high fundamental frequencies, it is critical that the load capacitance is accurately specified, usually in the range of 12pF to 30pF.

In parallel mode operation, the crystal will operate with an impedance in which the inductive reactance is dominant. In this condition, the shunt capacitance is the principal external influence in determining the operating frequency of the oscillator.

Part of the actual load capacitance consists of circuit and stray capacitances. These factors should be carefully evaluated to ensure that the actual load capacitance presented to the crystal is identical to the design value of load capacitance for which it is to operate.

Like capacitors, quartz crystals exhibit an equivalent series resistance (ESR). This is the series resistance value that the crystal exhibits when resonant. A crystal with low ESR is easier to excite into oscillation and delivers a cleaner, stronger

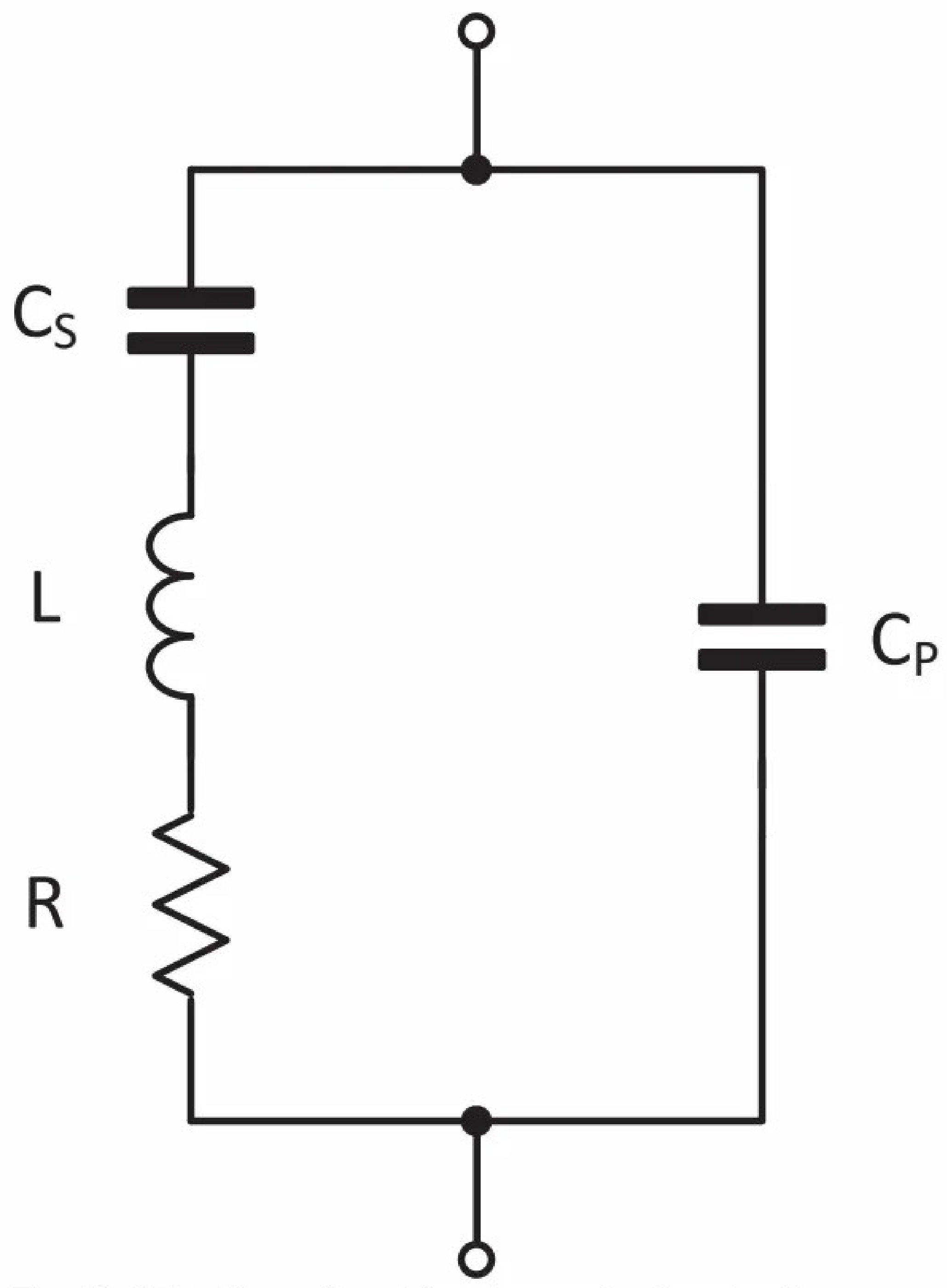


Fig.2.11: the electrical equivalent of a quartz crystal. There are two very close resonant frequencies: series and parallel.

output, so it's described as being more "active".

Consequently, quartz crystals with relatively low ESR values will require lower values of drive from the oscillator circuit to make the crystal resonate. Low-ESR crystals are therefore essential for low-voltage/low-current oscillators.

By introducing a combination of series and/or parallel reactance (both inductive and capacitive), the oscillation frequency of a crystal oscillator can be 'pulled' away from its nominal value by as much as ±0.5% of the nominal frequency (not necessarily symmetrically!).

While this is normally an undesirable effect, it can occasionally be useful in providing a way to vary the operating frequency of a crystal oscillator over a narrow range. Some radio manufacturers have exploited this variable crystal oscillator (VXO) technique in incremental and fine-tuning controls.

The frequency stability of a VXO arrangement is often significantly impaired at the edges of the 'pulling' range. Frequency stability can be preserved by keeping the excursion within reasonable limits.

Crystal filters

While individual quartz crystals are used as the frequency-determining element in crystal-controlled oscillators, multiple crystals can be used to produce highly selective filters.

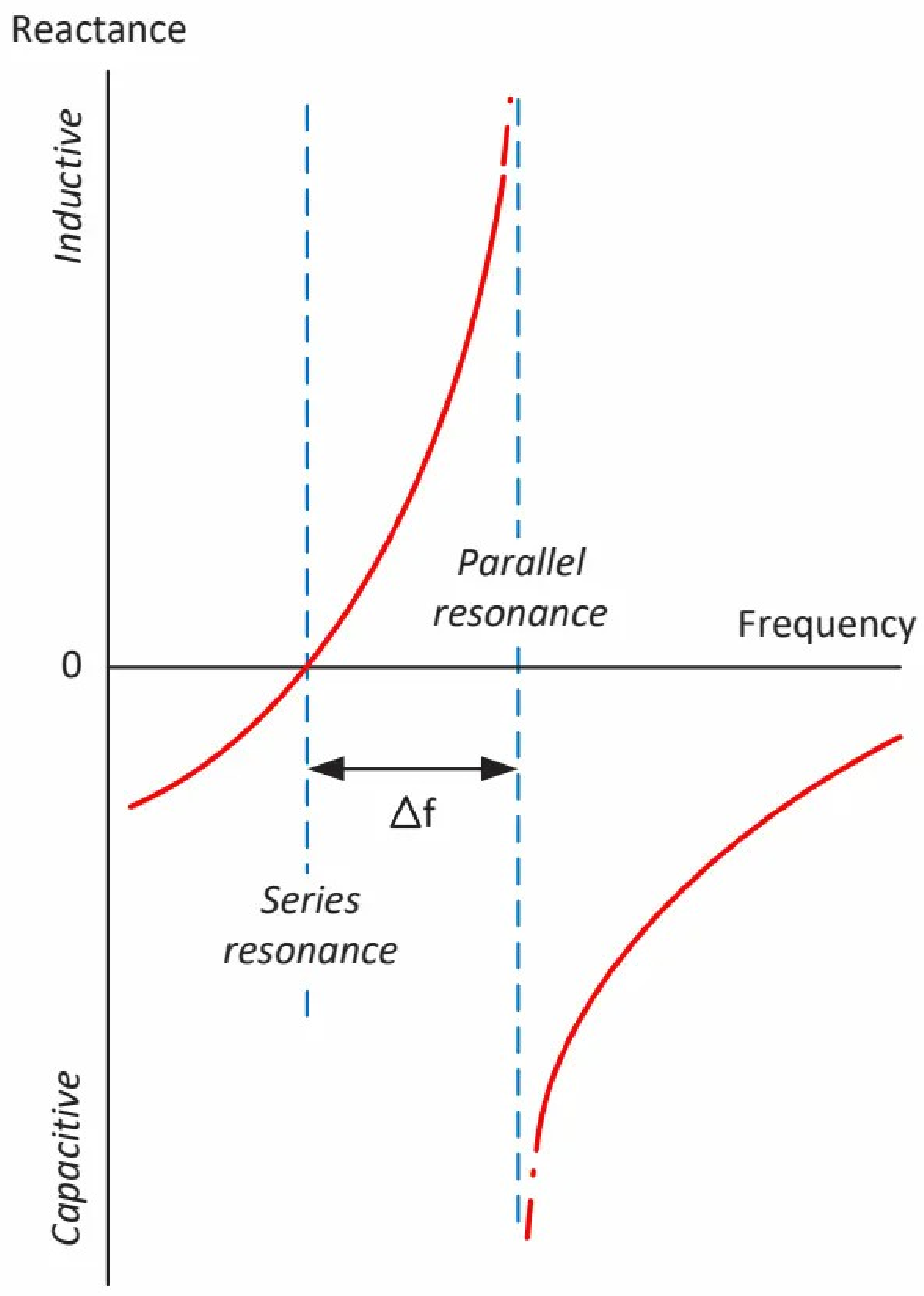
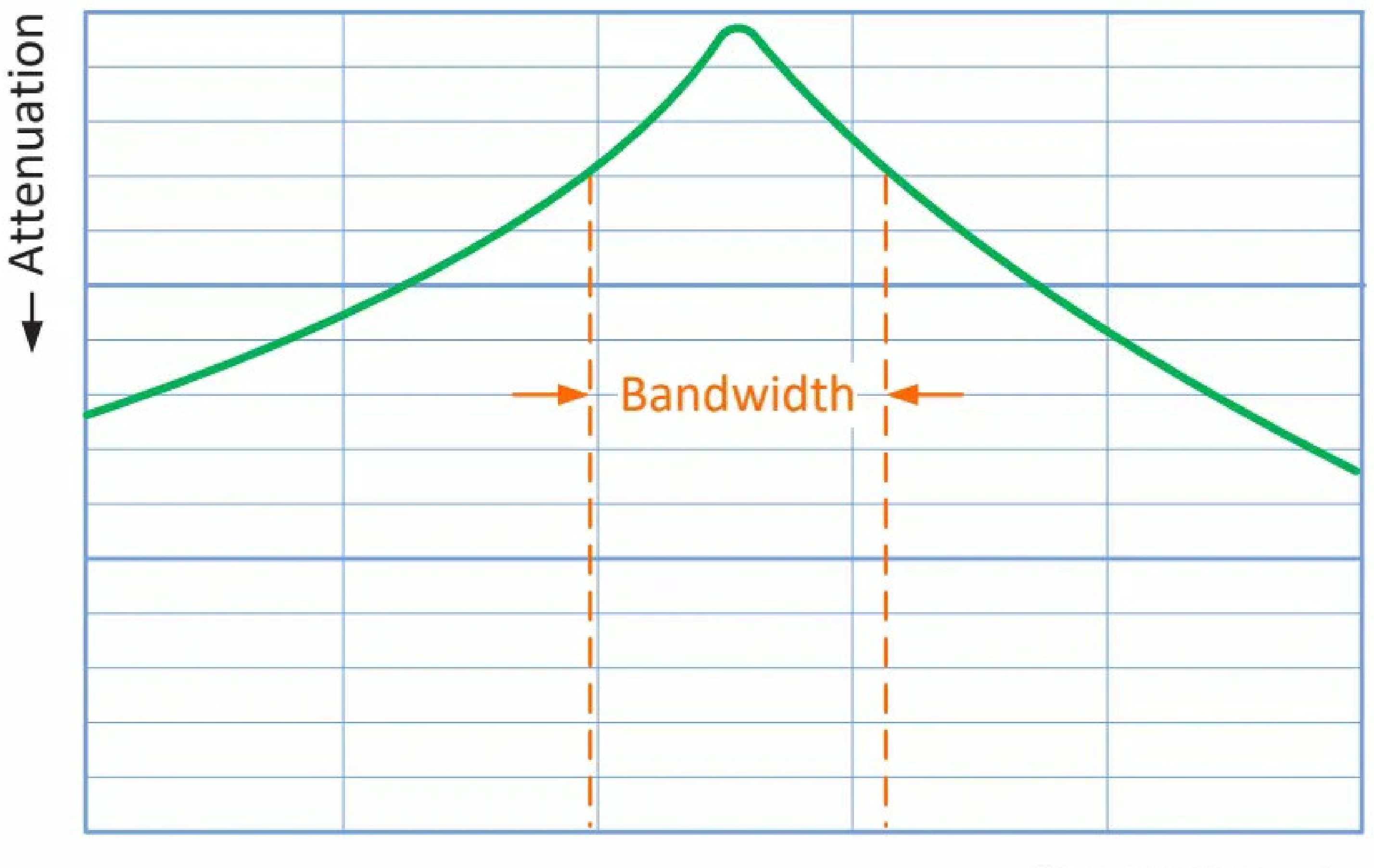


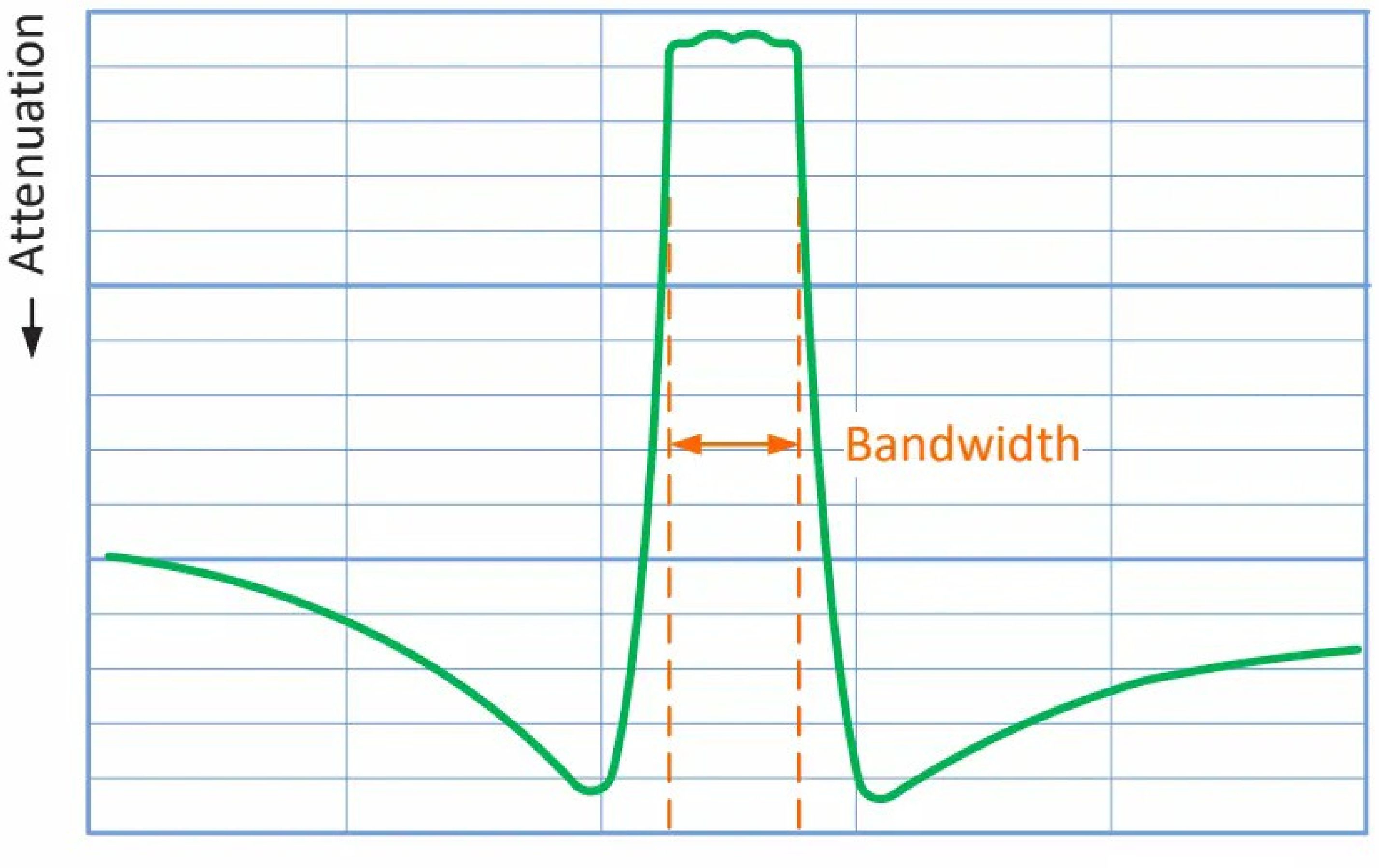
Fig.2.12: the frequency characteristic of a quartz crystal.

By careful selection of resonant frequencies, it is possible to produce a network of crystals with near-perfect filter characteristics: a very low loss in the passband, coupled with a very high attenuation in the stopband.

Fig.2.13 shows how the frequency response of a crystal filter compares with that of a conventional L-C circuit filter. Note that the response curve of the crystal filter shows a



(a) L-C filter



(b) Crystal filter

Fig.2.13: a comparison of L-C and crystal filters.

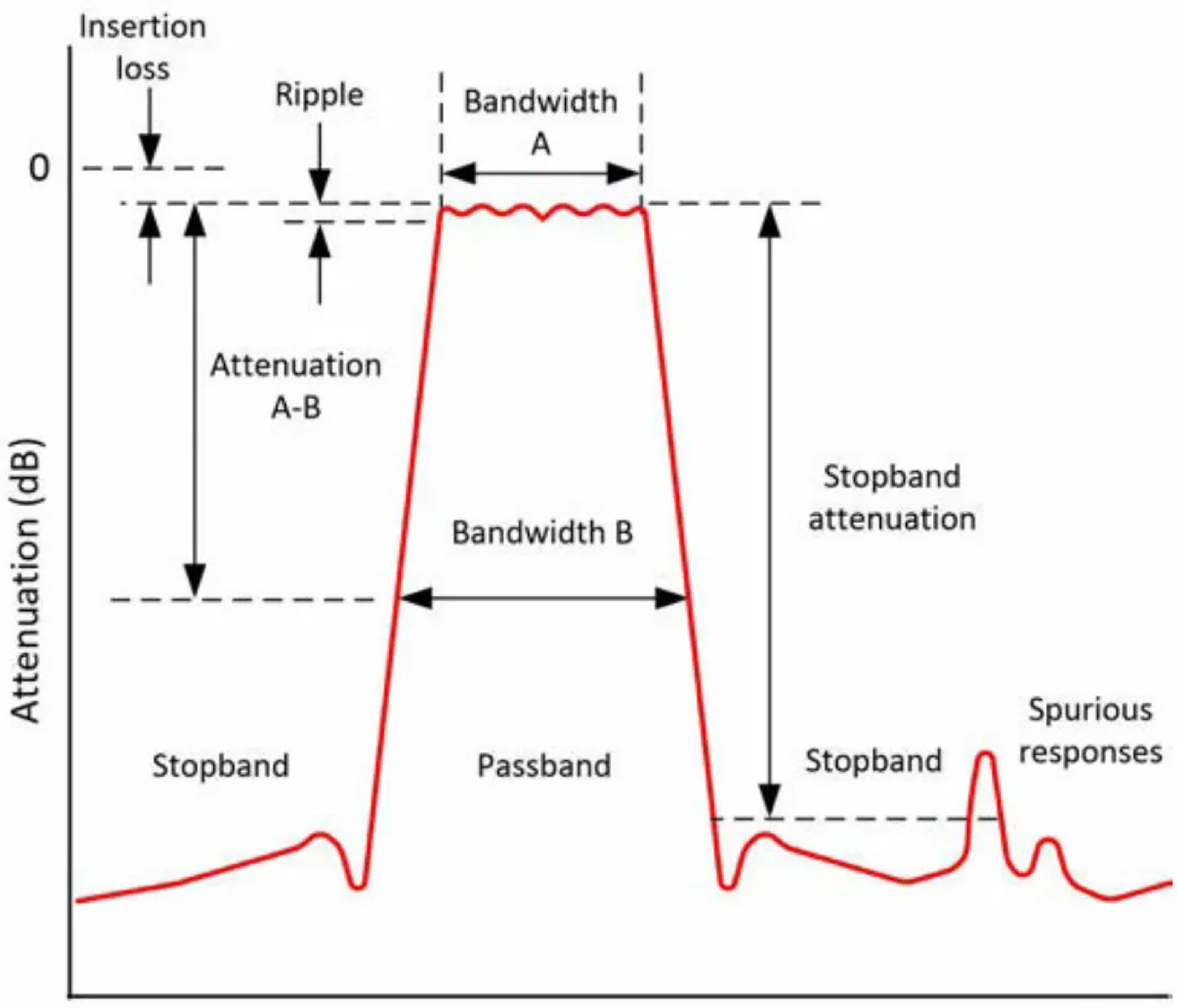


Fig.2.14: crystal filter frequency response.

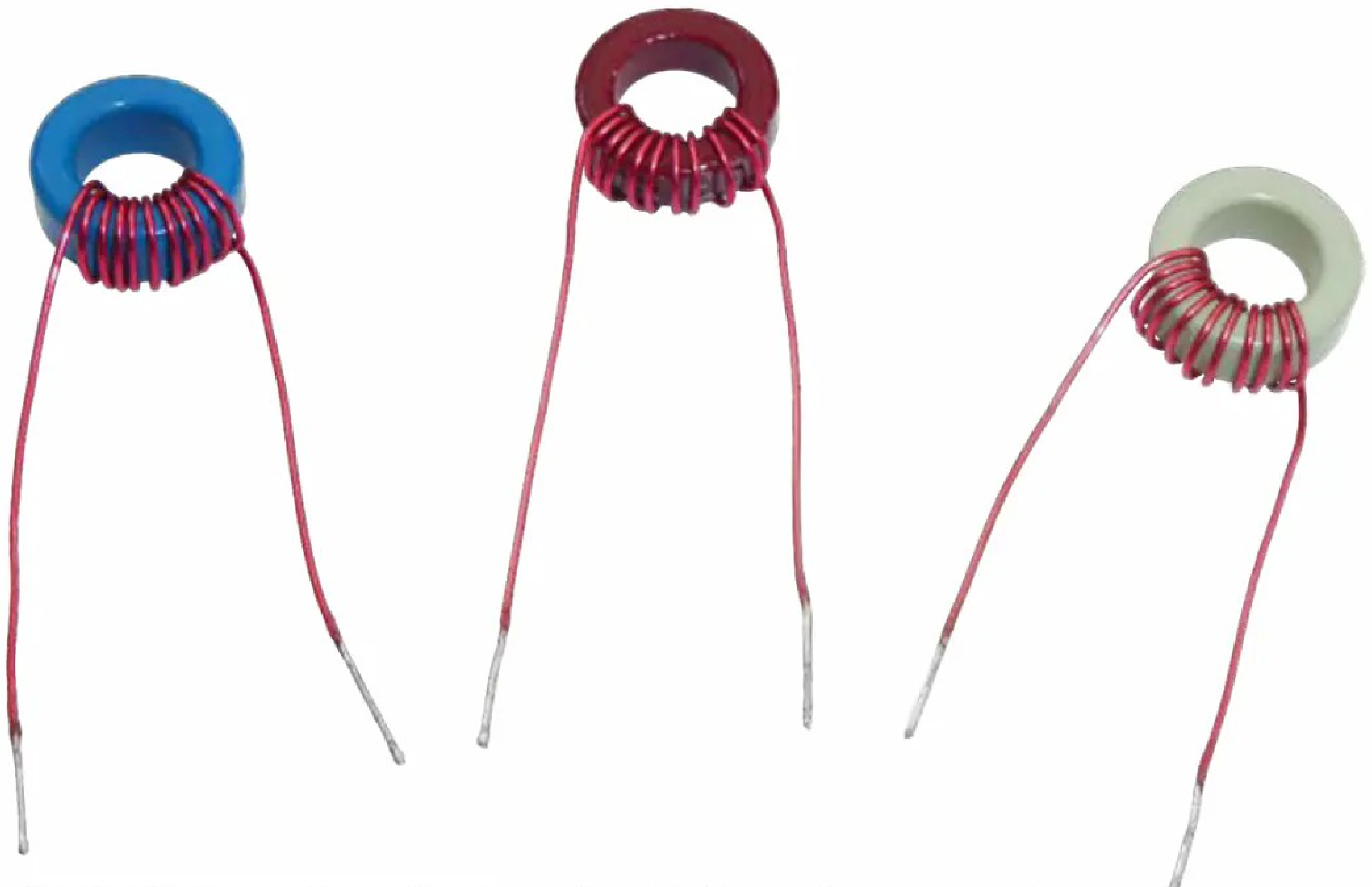


Fig.2.19: these three ten-turn toroidal inductors have very different characteristics due to the different core materials. They exhibit inductances (left to right) of 265µH, 9µH, and 2.5µH. Only the right-most core is suitable for high-Q applications.

flatter top and steeper sides. This characteristic is essential for effective rejection of signals on adjacent frequencies.

A typical crystal filter frequency response (showing the passband and stopband regions) is shown in Fig.2.14. The stopband attenuation for this filter is typically more than 60dB with a passband ripple under 3dB.

The shape factor for the filter is usually specified in terms of the passband bandwidth (A) and the attenuation in the stopband, bandwidth (B).

Thus, for a given value of attenuation, Shape factor = (Bandwidth B) ÷ (Bandwidth A). Typical values of shape factor range from about 1.5 to 3.

The spurious responses (see Fig.2.14) associated with quartz crystal filters can be made less significant by using conventional tuned-circuit matching transformers. A typical 9MHz crystal filter for use in an HF communications receiver might have the following specifications:

- **Centre frequency**: 9MHz
- **Passband bandwidth**: 4kHz @ -6dB
- **Stopband bandwidth**: 12kHz @ -60dB
- **Maximum ripple**: 2dB
- **Maximum insertion loss**: 3dB
- **Input/output impedance**: 1.5kΩ
- **60dB shape factor**: 3

Unfortunately, good-quality crystal filters can be extremely expensive. For less critical applications, ceramic filters make a more cost-effective alternative. These are available in several standard frequencies, including 455kHz and 10.7MHz. We will be exploring their use later in this series.

RF inductors and transformers

RF inductors and transformers, like those shown in Fig.2.15, are frequently needed in RF circuits. For convenience, we've mainly used standard fixed-value inductors in our Hands-On projects, but you may occasionally need a specific value inductance that's unavailable as a standard component or that needs to be made adjustable.

In such cases, winding your own coil on a small former can be the best solution. The problem of determining the dimensions and current number of turns can be solved in various ways. However, coil winding is not a precise science,

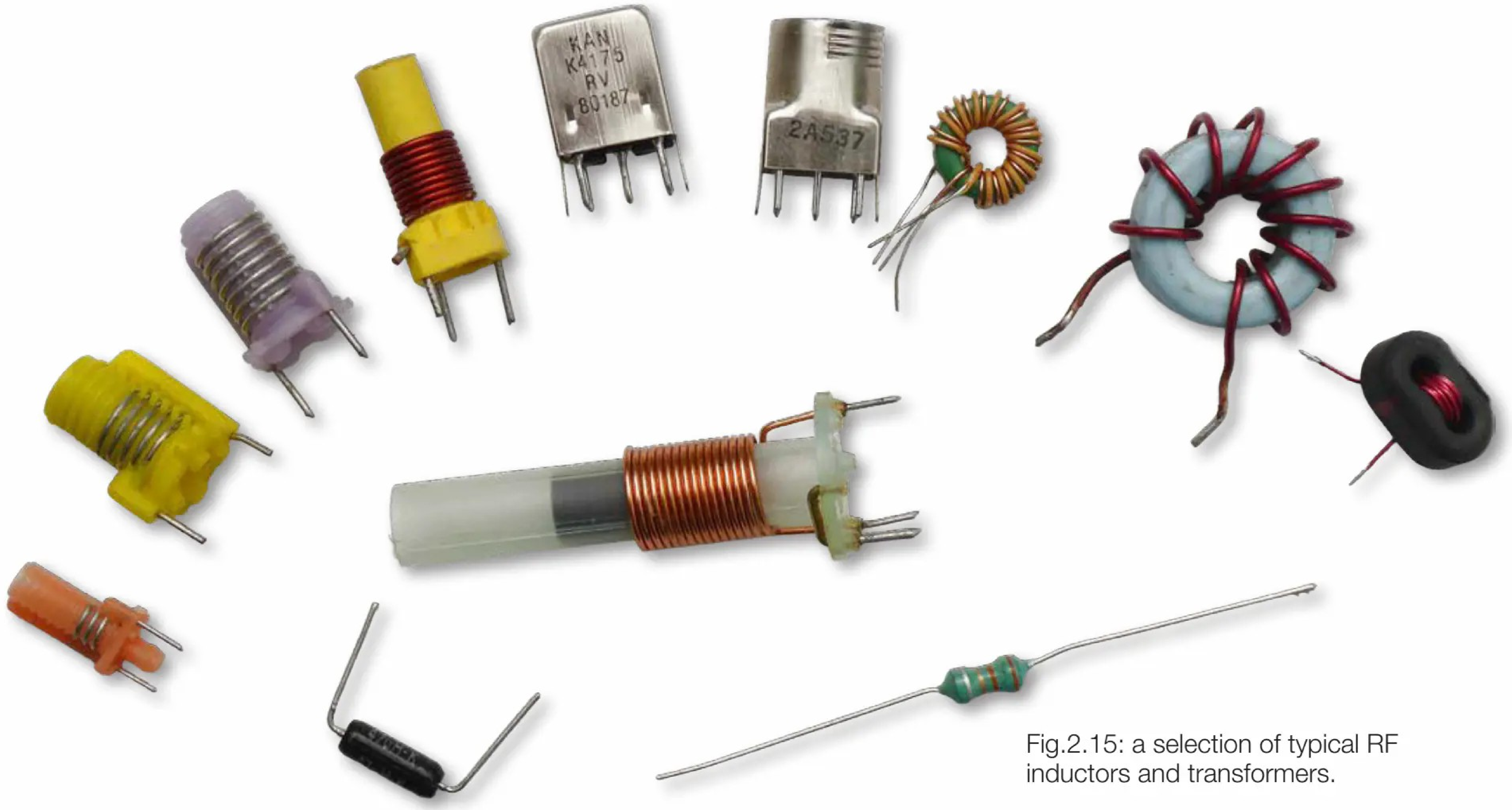


Fig.2.15: a selection of typical RF inductors and transformers.

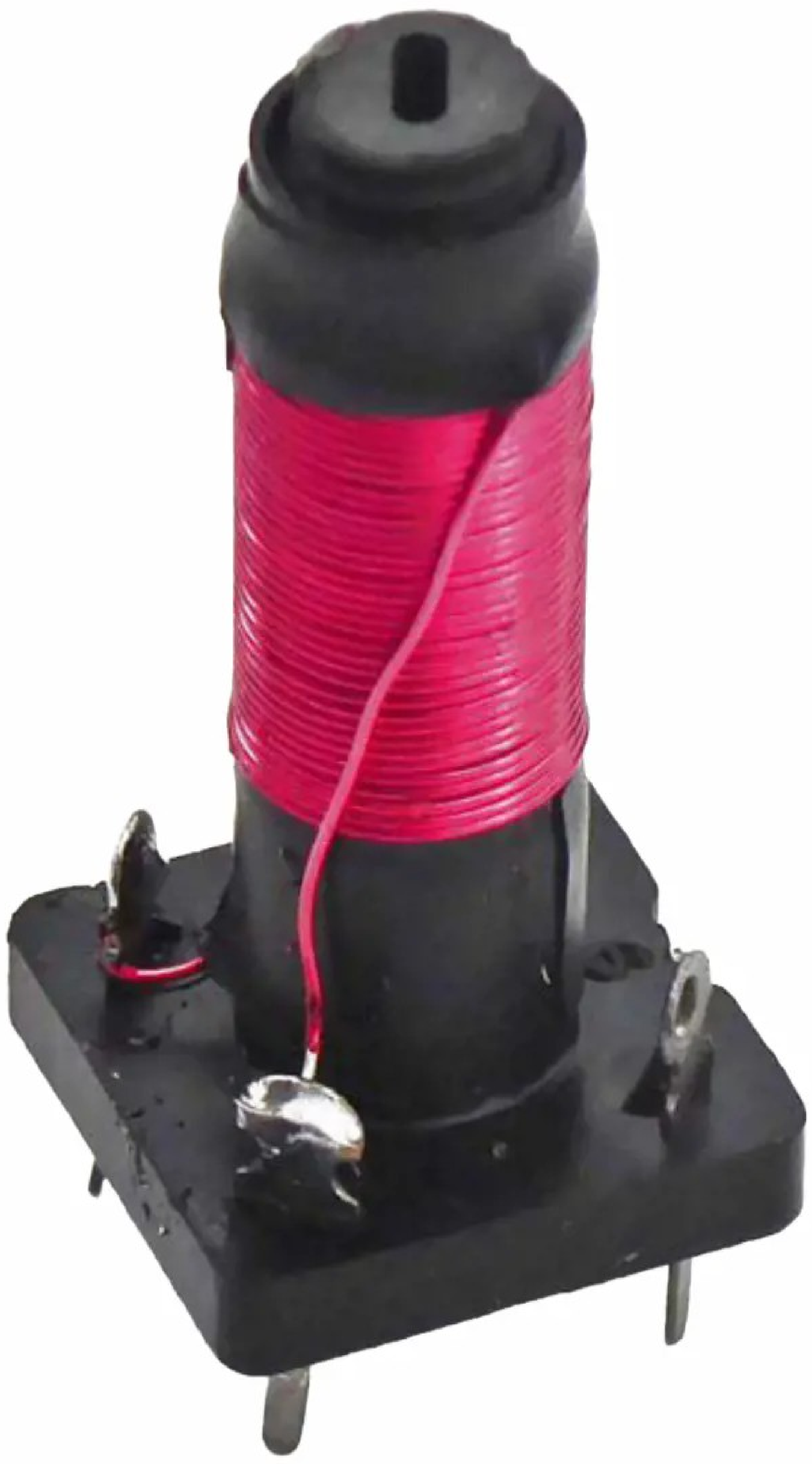


Fig.2.17: the 9µH adjustable inductor.

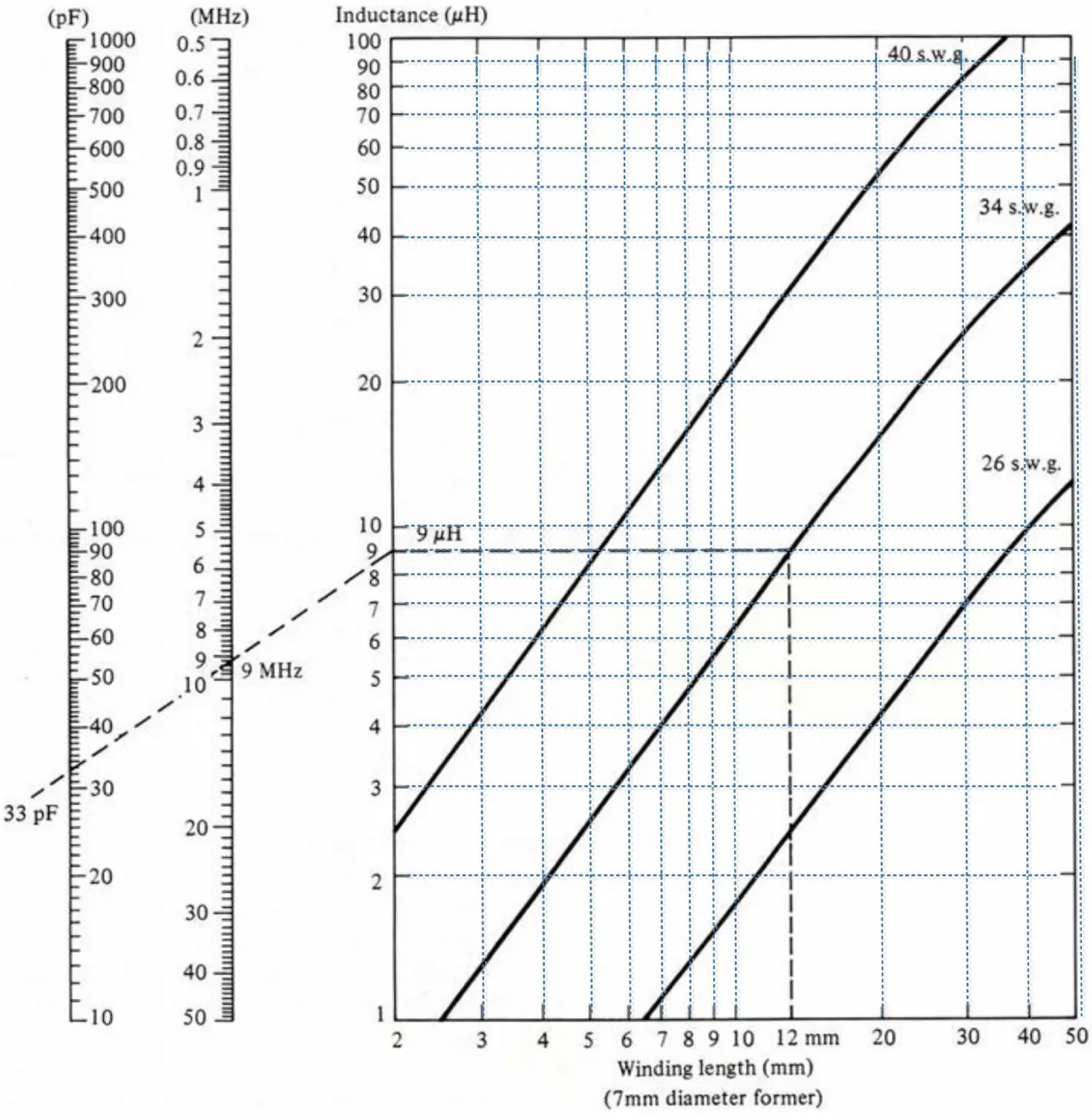


Fig.2.16: a coil winding chart for 7mm formers.

so experimentation will often be required.

The coil winding chart shown in Fig.2.16 provides a quick rule-of-thumb method of determining the length of a coil winding required on a typical 7mm former. When supplied in insulated enamelled copper form, the three wire gauges quoted (26 SWG, 34 SWG and 40 SWG) will have typical diameters of 0.5mm, 0.3mm and 0.2mm, respectively.

The example in Fig.2.16 shows how a 9µH inductor designed for operation at 9MHz and tuned by a 33pF capacitor will need to span around 12mm of a 7mm former when wound with 34 SWG (0.3mm diameter) enamelled copper wire (ECW). The resulting inductor is shown in Fig.2.17.

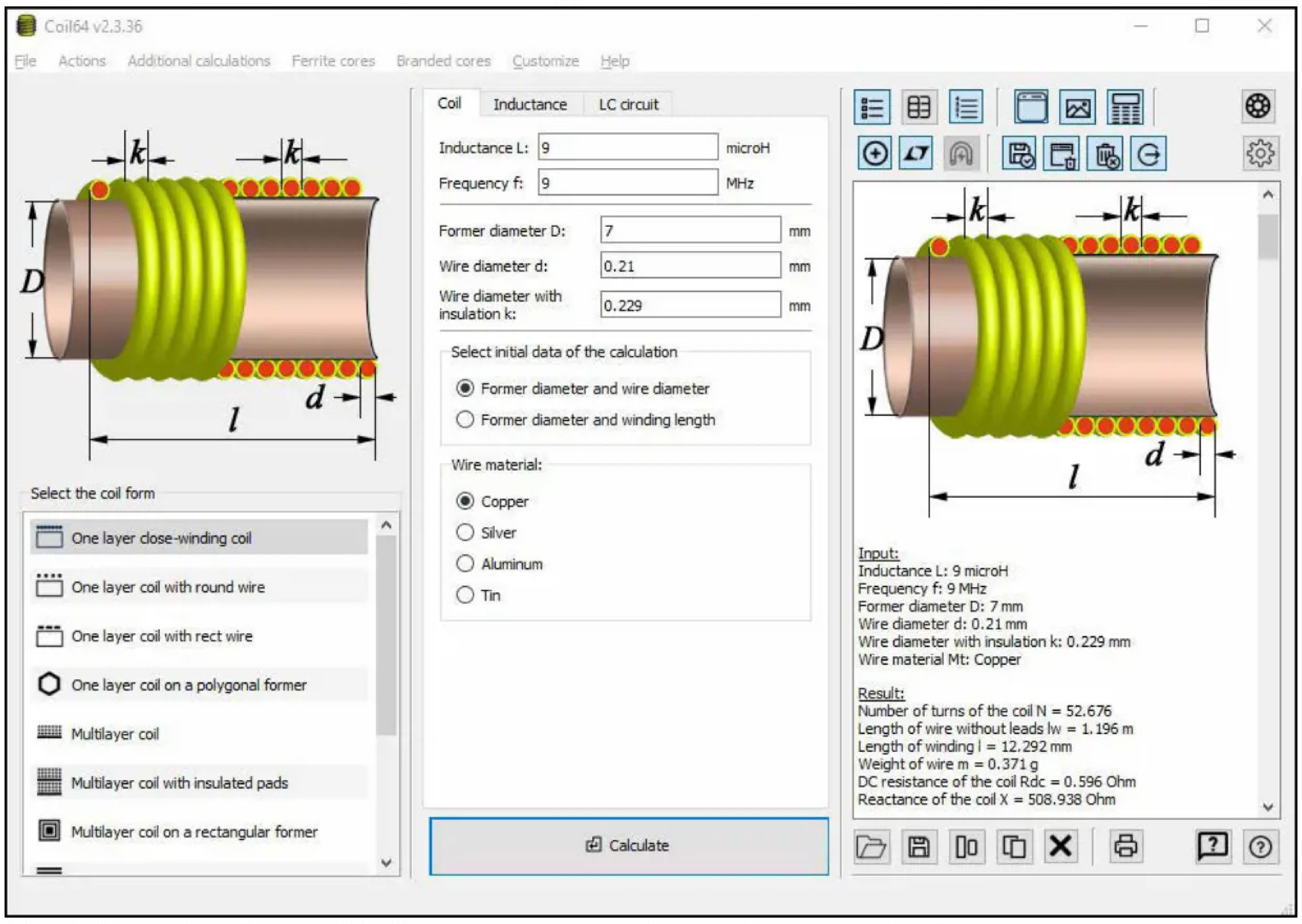
When using an adjustable ferrite core, the winding length (and number of turns) will typically need to be reduced by about 30%. The component shown in Fig.2.17 exhibits an inductance of between 7µH and 22µH over the adjustment range of the ferrite core, corresponding to resonant frequencies of 6-10MHz when tuned with 33pF.

Coil64 (see Fig.2.18) offers a very neat solution for designing many types of inductors. It is a 64-bit, cross-platform, open-source application licensed under GPLv3. It runs on desktop Linux, macOS (64-bit), and Windows (32/64-bit) systems, and the application can be freely downloaded from **https://coil32.net**

Toroidal ferrite cores are widely used in RF circuits for low-Q inductors, filters and impedance matching. Their core material determines critical characteristics such as inductance value, Q-factor and suitability for specific applications.

One significant advantage of, and reason for using, toroidal cores is that their magnetic fields are mostly contained within and immediately around the core, so they interact less with other nearby coils compared to simple cylindrical forms.

As illustrated in Fig.2.19, toroidal cores that have identical windings can yield vastly different inductance values depending on the material. One may produce high inductance (eg, 265µH), another moderate (eg, 9µH), and a third, low (eg, 2.5µH).



Inductance L: 9 microH
Frequency f: 9 MHz
Former diameter D: 7 mm
Wire diameter d: 0.21 mm
Wire diameter with insulation k: 0.229 mm

Select initial data of the calculation
● Former diameter and wire diameter
○ Former diameter and winding length

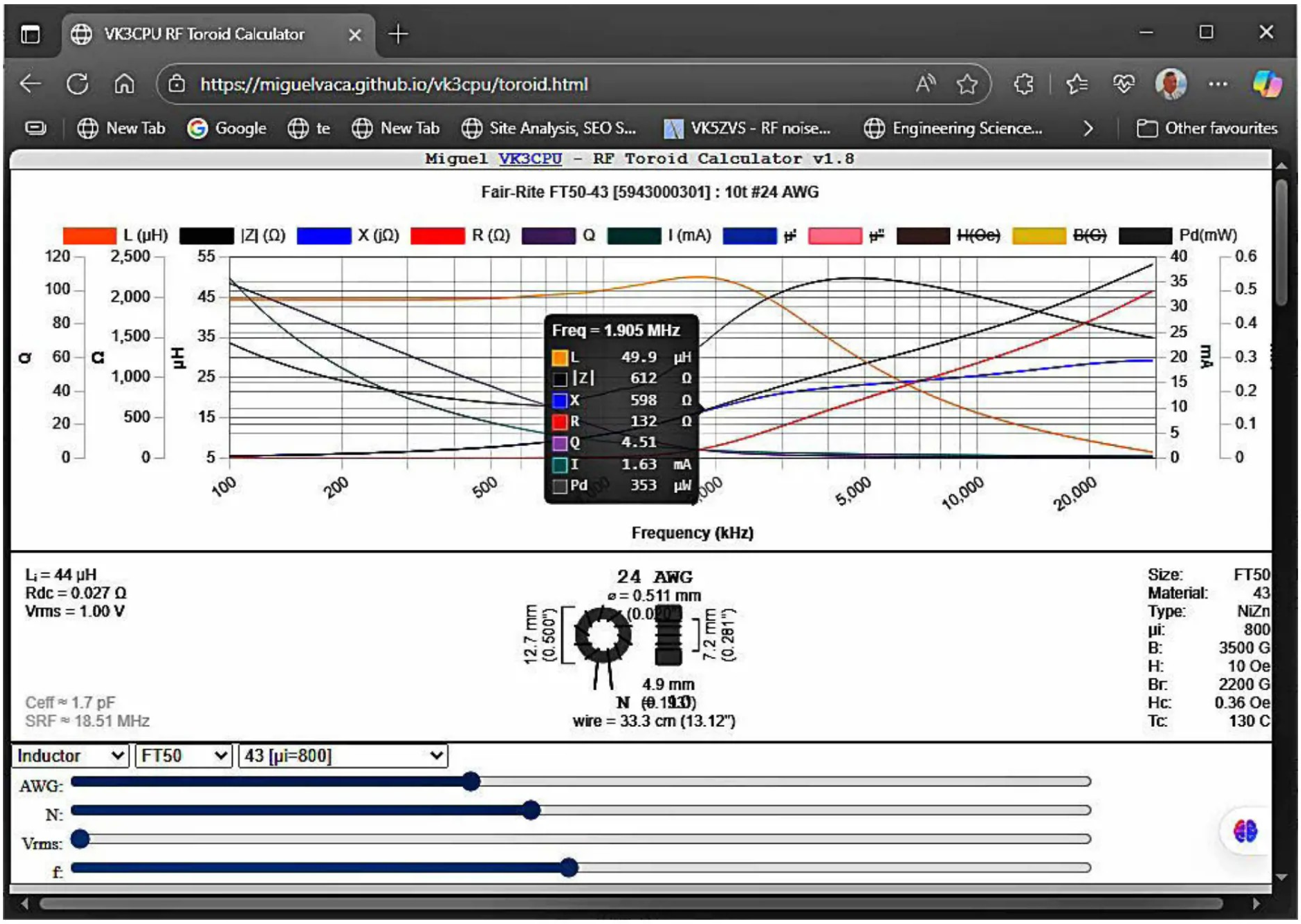
Wire material:
● Copper
○ Silver
○ Aluminum
○ Tin

Select the coil form
One layer close-winding coil
One layer coil with round wire
One layer coil with rect wire
One layer coil on a polygonal former
Multilayer coil
Multilayer coil with insulated pads
Multilayer coil on a rectangular former

Input:
Inductance L: 9 microH
Frequency f: 9 MHz
Former diameter D: 7 mm
Wire diameter d: 0.21 mm
Wire diameter with insulation k: 0.229 mm
Wire material Mt: Copper

Result:
Number of turns of the coil N = 52.676
Length of wire without leads lw = 1.196 m
Length of winding l = 12.292 mm
Weight of wire m = 0.371 g
DC resistance of the coil Rdc = 0.596 Ohm
Reactance of the coil X = 508.938 Ohm

Fig.2.18: using the Coil64 freeware application for inductor design.


VK3CPU RF Toroid Calculator
New Tab
Google
te
New Tab
Site Analysis, SEO S...
VK5ZVS - RF noise...
Engineering Science...
Other favourites
L (µH)
|Z| (Ω)
X (jΩ)
R (Ω)
Q
I (mA)
µ'
µ"
~~H(Oe)~~
~~B(G)~~
Pd(mW)
Q
Ω
µH
mA
120
100
80
60
40
20
0
2,500
2,000
1,500
1,000
500
0
55
45
35
25
15
5
40
35
30
25
20
15
10
5
0
0.6
0.5
0.4
0.3
0.2
0.1
0
100
200
500
5,000
10,000
20,000
Frequency (kHz)
12.7 mm
(0.500")
7.2 mm
(0.281")
4.9 mm
Inductor
FT50
43 [µi=800]
AWG:
N:
Vrms:
f:

https://miguelvaca.github.io/vk3cpu/toroid.html

Miguel VK3CPU - RF Toroid Calculator v1.8

Fair-Rite FT50-43 [5943000301] : 10t #24 AWG

Freq = 1.905 MHz
L 49.9 µH
|Z| 612 Ω
X 598 Ω
R 132 Ω
Q 4.51
I 1.63 mA
Pd 353 µW

L_i = 44 µH
Rdc = 0.027 Ω
Vrms = 1.00 V

Ceff ≈ 1.7 pF
SRF ≈ 18.51 MHz

24 AWG
⌀ = 0.511 mm
wire = 33.3 cm (13.12")

Size: FT50
Material: 43
Type: NiZn
µi: 800
B: 3500 G
H: 10 Oe
Br: 2200 G
Hc: 0.36 Oe
Tc: 130 C

Fig.2.20: the RF Toroidal Core application, showing the frequency characteristic of a 45µH inductor from 100kHz to 20MHz (the orange trace; note the legend above the plot). There is a progressive reduction in inductance above 2MHz.

Only certain materials are suitable for high-Q applications such as precise RF filtering and tuning. Therefore, selecting the appropriate core material is essential—some are optimised for low-Q filtering and matching, while others are designed for high-Q, high-frequency performance. Careful attention to core material properties is needed to ensure reliable and effective RF performance.

The design of toroidal core inductors is greatly simplified with the use of software. Fig.2.20 shows Miguel Vaca's excellent online toroidal calculator. You must specify the type and material of the core before experimenting with other parameters, such as wire gauge, turns and frequency. RF Toroid Calculator is available at: https://miguelvaca.github.io/vk3cpu/toroid.html

With toroidal core inductors, it is important to be aware that inductance increases with frequency and DC bias. Power loss also increases with frequency and flux density, and this can be an important consideration when toroidal cores are to be used in power amplifiers.

Finally, toroidal cores are often supplied with blue, grey, yellow, silver, red or green markings, but there appears to be no universally accepted standard for colour coding. One reliable supplier of cores, DX Engineering, uses the coding shown in Table 2.1.

RF semiconductors

The choice of semiconductor devices can be crucial in many RF applications. Devices designed primarily for low-frequency applications (below 3MHz) will usually not perform well at HF, VHF and UHF. Selecting the right device for the task can be crucial, so here's a shortlist of selected RF semiconductors, including many of those that feature in our Hands-On projects.

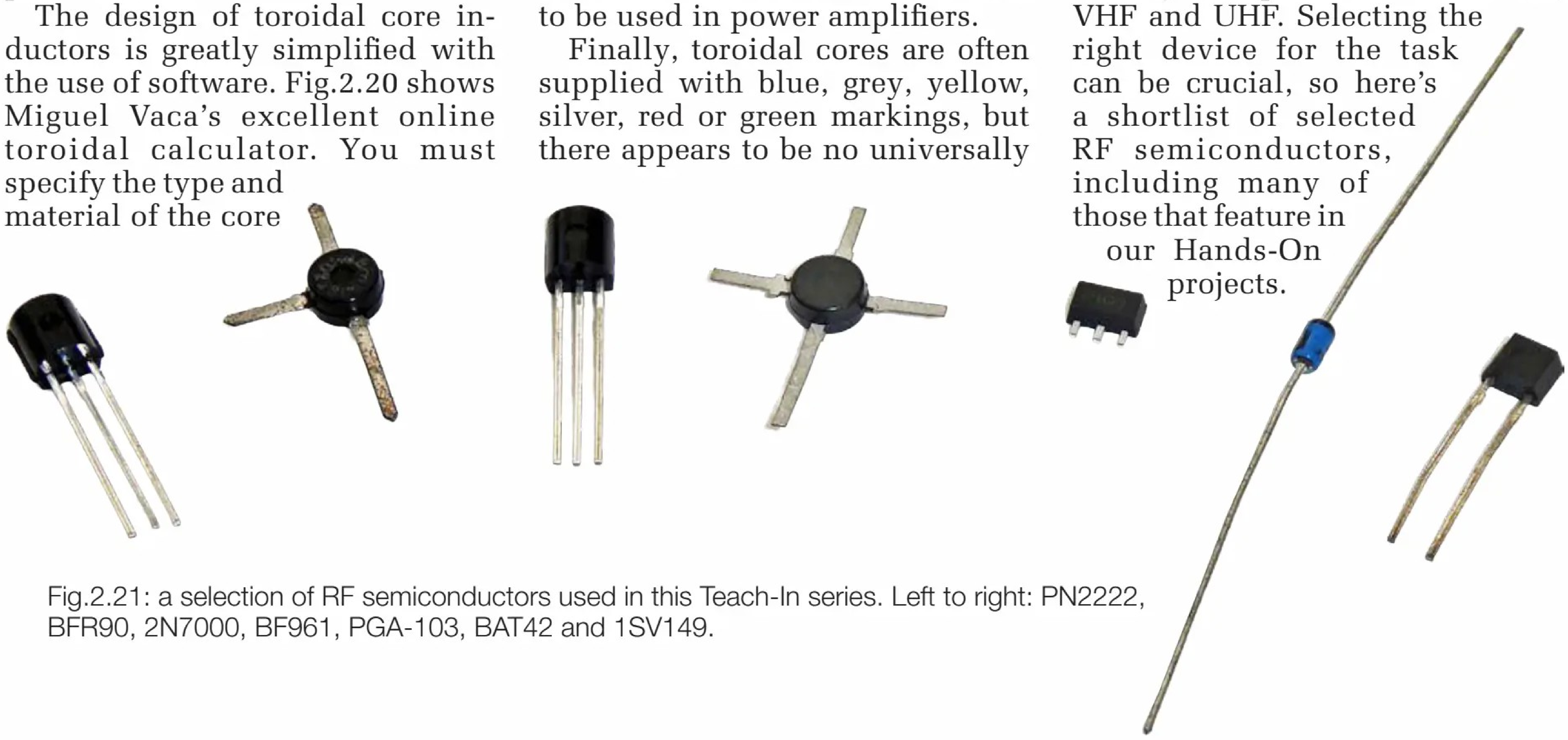


Fig.2.21: a selection of RF semiconductors used in this Teach-In series. Left to right: PN2222, BFR90, 2N7000, BF961, PGA-103, BAT42 and 1SV149.

Table 2.1 – Colour coding of toroidal cores supplied by DX Engineering

Colour marking	Core material	Frequency range	Application
Blue	31	1-300MHz	HF and VHF
Silver	43	25-300MHz	VHF
Orange	52	200-1000MHz	VHF and UHF
Red	61	200-2000MHz	VHF and UHF
Green	75	0.15-30MHz	MF and HF
Pink	77	0.1-50MHz	LF, MF and HF

2N2222 NPN silicon transistor

The 2N2222 is a popular NPN bipolar junction transistor (BJT) used in RF and switching circuits. It offers moderate gain, fast switching speed and the ability to handle relatively high currents and voltages. Typically, the 2N2222 can handle collector currents up to 600mA and voltages up to 40V, making it ideal for driving inductive loads.

The TO-18 metal case variant is still available, but the low-cost TO-92 packaged version is available as the PN2222. The device has a typical transition frequency (f_T) of around 250MHz, so is suitable for operation up to and beyond 100MHz. This makes it ideal for non-critical HF and low-VHF applications.

BFR90 NPN silicon transistor

The BFR90 is a high-frequency NPN silicon BJT specifically designed for VHF RF applications. It has a low noise figure, together with excellent gain, making it ideal for use in amplifiers, oscillators and mixers operating up to 450MHz. The BFR90 comes in a TO-50 package, visible in Fig.2.21, with a maximum collector current of around 25mA and a maximum collector-emitter voltage of 15V.

BF245C N-channel JFET

The BF245 is an N-channel junction field-effect transistor (JFET) commonly used in a wide range of high-frequency, low-noise applications, including RF amplifiers, oscillators, and mixers operating up to 150MHz. Variants like the BF245A, BF245B and BF245C have slightly different characteristics and may require different circuit parameters. The device is supplied in a TO-92 package.

2N7000 N-channel Mosfet

The 2N7000 is a widely used N-channel enhancement-mode Mosfet offering fast switching speed, relatively low on-resistance (typically less than 5Ω) and a very high input impedance (in common-source mode).

The device is supplied in a TO-92 plastic package and is capable of handling drain currents up to 200mA and drain-source voltages of up to 60V. RF applications include oscillators, buffer amplifiers, and drivers in low-power transmitters. The 2N7002 is a similar device in an SMD (SOT-23) package and comes in a few different variants (2N7002P, 2N7002K etc), although some of those are not as good for RF applications.

BF961 dual gate Mosfet

The BF961 is a dual-gate N-channel depletion-mode Mosfet, well-regarded for its performance in high-frequency RF applications. The device is often found in the front-end stages of VHF and UHF receivers, where its excellent gain, low noise figure and high input impedance are particularly useful.

The dual-gate configuration permits effective gain control and improved signal selectivity, making it ideal for use in mixers, amplifiers, and RF AGC (automatic gain control) circuits. The device is supplied in a four-lead TO-5 package, seen in the middle of Fig.2.21. A depletion-mode Mosfet is on until biased off, while the more common enhancement-mode type is off until biased on.

PGA-103 low-noise amplifier (LNA)

Like the immensely popular SPF5189Z LNA, the PGA-103 is a high-gain, ultra-low-noise RF amplifier that offers outstanding performance across a range of frequencies from 560MHz to 3GHz. The device boasts a typical noise figure of around 0.5dB at 150MHz with a gain of 24-25dB. This makes it suitable for front-end RF amplifiers, satellite receivers and RF measuring instruments.

The device has straightforward biasing requirements and exhibits excellent linearity, with a high third-order intercept point (IP3) of around 36dB, indicating excellent linearity. The PGA-103 operates from a 5V DC supply and comes in an SOT-89 surface-mount package.

The PGA-103 distinguishes itself from standard general-purpose RF amplifiers like the BFR90 or BF961 previously described by offering a much lower noise figure and higher gain, particularly at VHF and UHF frequencies.

While the BFR90 is suitable for small-signal amplification, and the BF961 excels as a dual-gate Mosfet mixer or AGC amplifier, devices like the PGA-103 or SPF5189 are better when low noise, high gain, and exceptional frequency response is required.

BAT42 schottky diode

The BAT42 is a small-signal schottky fast-switching diode that exhibits a low forward voltage drop of around 0.3V. The device is widely used in applications such as high-frequency rectification, clamping and protection circuits, especially where efficiency and fast response are critical. Its low capacitance and high reliability make it well suited to a wide variety of RF, digital and signal processing tasks.

Schottky diodes conduct using only majority carriers, so they don't suffer from the minority-carrier charge storage of standard PN diodes. This eliminates the reverse-recovery delay when switching from forward to reverse bias.

1SV149 variable capacitance diode

The 1SV149 is a variable capacitance diode, or varactor, designed

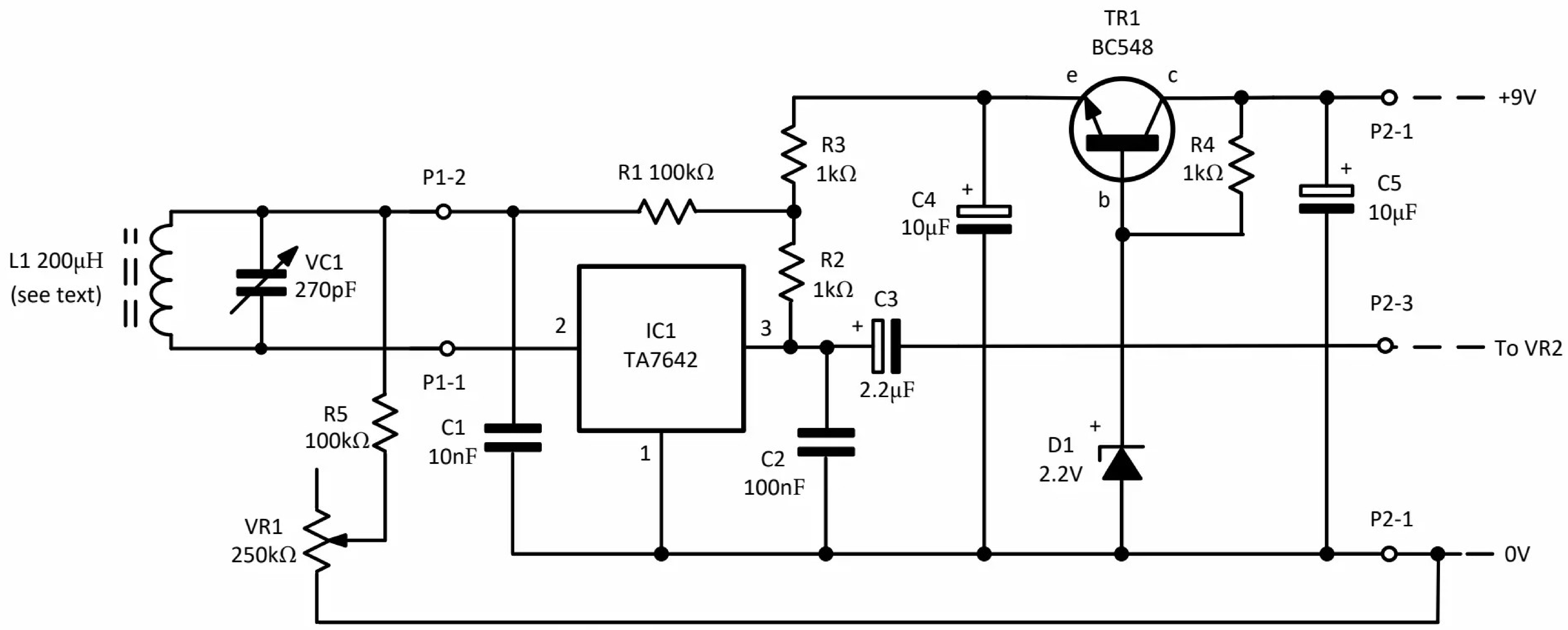

TR1
BC548
e
c
b
+9V
P2-1
R4
1kΩ
C5
10µF
R3
1kΩ
C4
10µF
R1 100kΩ
P1-2
L1 200µH
(see text)
VC1
270pF
R2
1kΩ
C3
2.2µF
IC1
TA7642
2
3
1
P2-3
To VR2
P1-1
R5
100kΩ
C1
10nF
C2
100nF
D1
2.2V
VR1
250kΩ
P2-1
0V

Fig.2.22: the RF module circuit for our new radio receiver.

for electronic tuning applications. It offers a wide capacitance range that changes proportionally with applied reverse voltage, making it suitable for voltage-controlled oscillators, frequency modulators and tuneable filters in a wide variety of RF circuits.

It offers low series resistance and stable performance across a wide range of frequencies, making it a preferred choice for both communications and general-purpose radio projects.

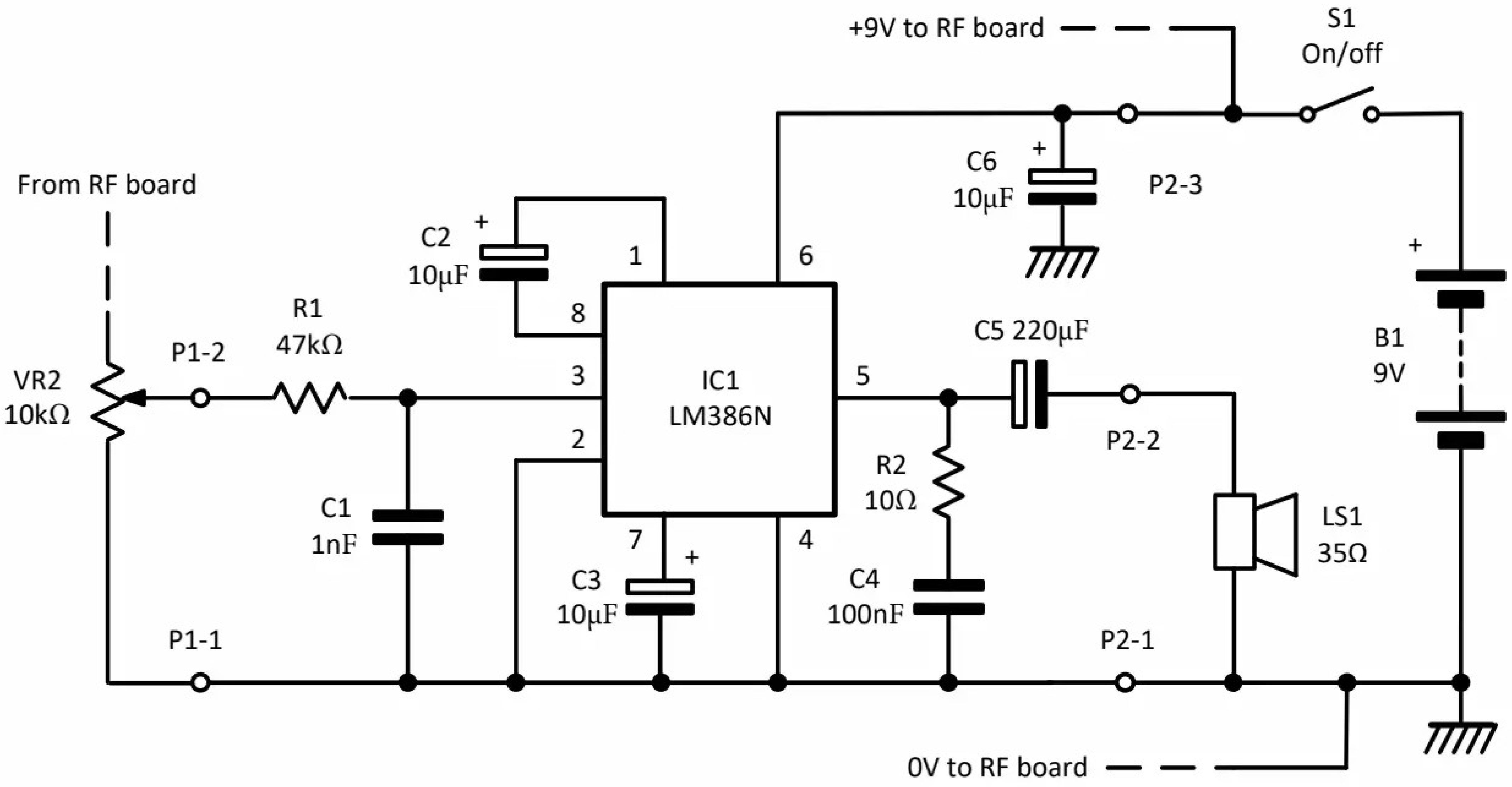
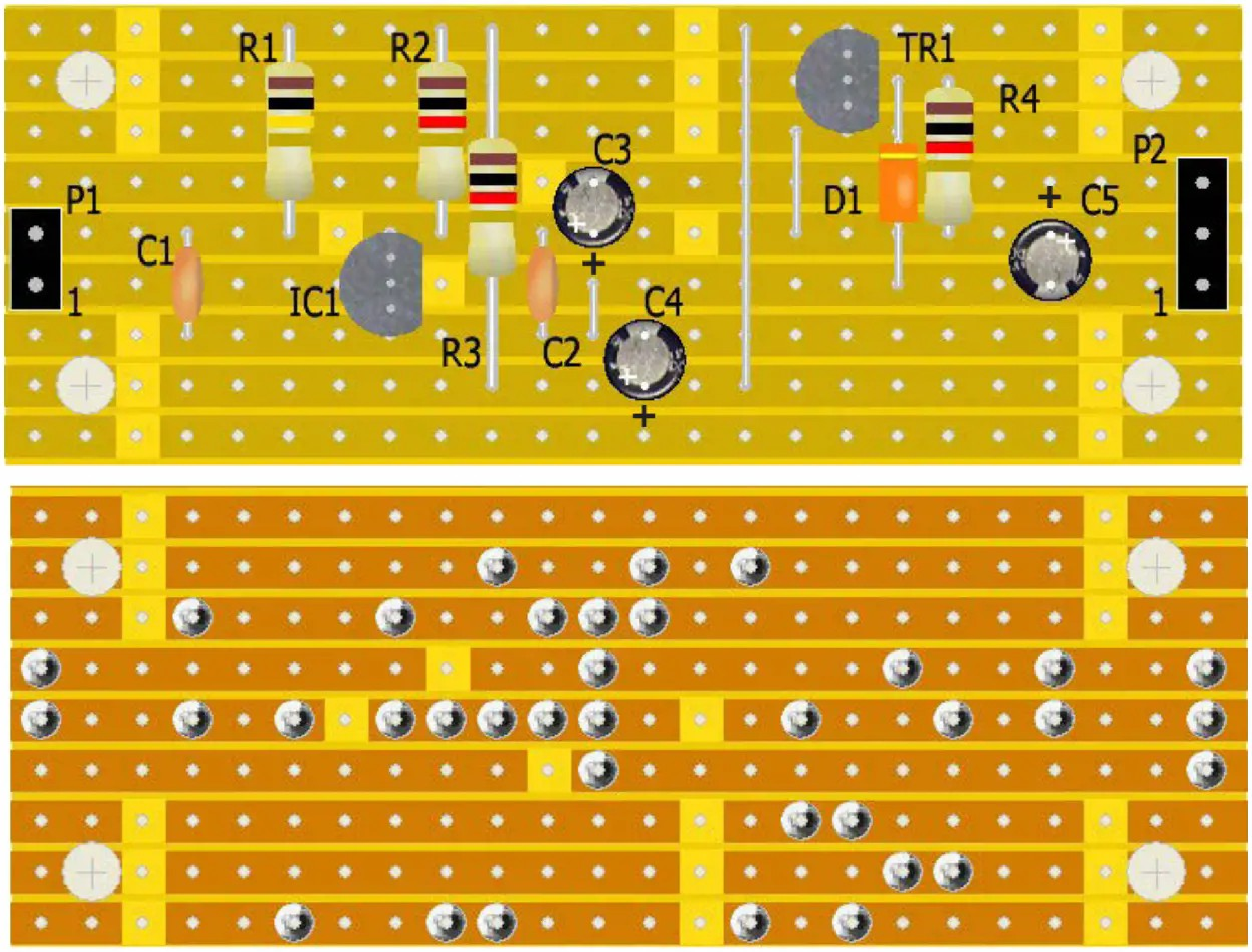

+9V to RF board
S1
On/off
From RF board
C6
10µF
P2-3
C2
10µF
1
6
8
R1
47kΩ
P1-2
VR2
10kΩ
3
IC1
LM386N
5
C5 220µF
B1
9V
2
R2
10Ω
P2-2
C1
1nF
7
4
LS1
35Ω
C3
10µF
C4
100nF
P1-1
P2-1
0V to RF board

Fig.2.23: the AF module circuit for our new radio receiver.


R1
R2
TR1
R4
C3
P2
P1
D1
+ C5
C1
IC1
1
1
C4
R3
C2

Figs.2.24 & 2.25: the component layout (top) and track view (below) for the RF module.

Hands-On: A portable AM receiver

This month's Hands-On project is a portable AM receiver, which builds on last month's project by offering an RF gain control and much improved audio output from a small loudspeaker. The receiver uses a 9V power supply provided by a battery of six AA 1.5V cells connected in series (you could use a smaller 9V PP3 battery if you prefer).

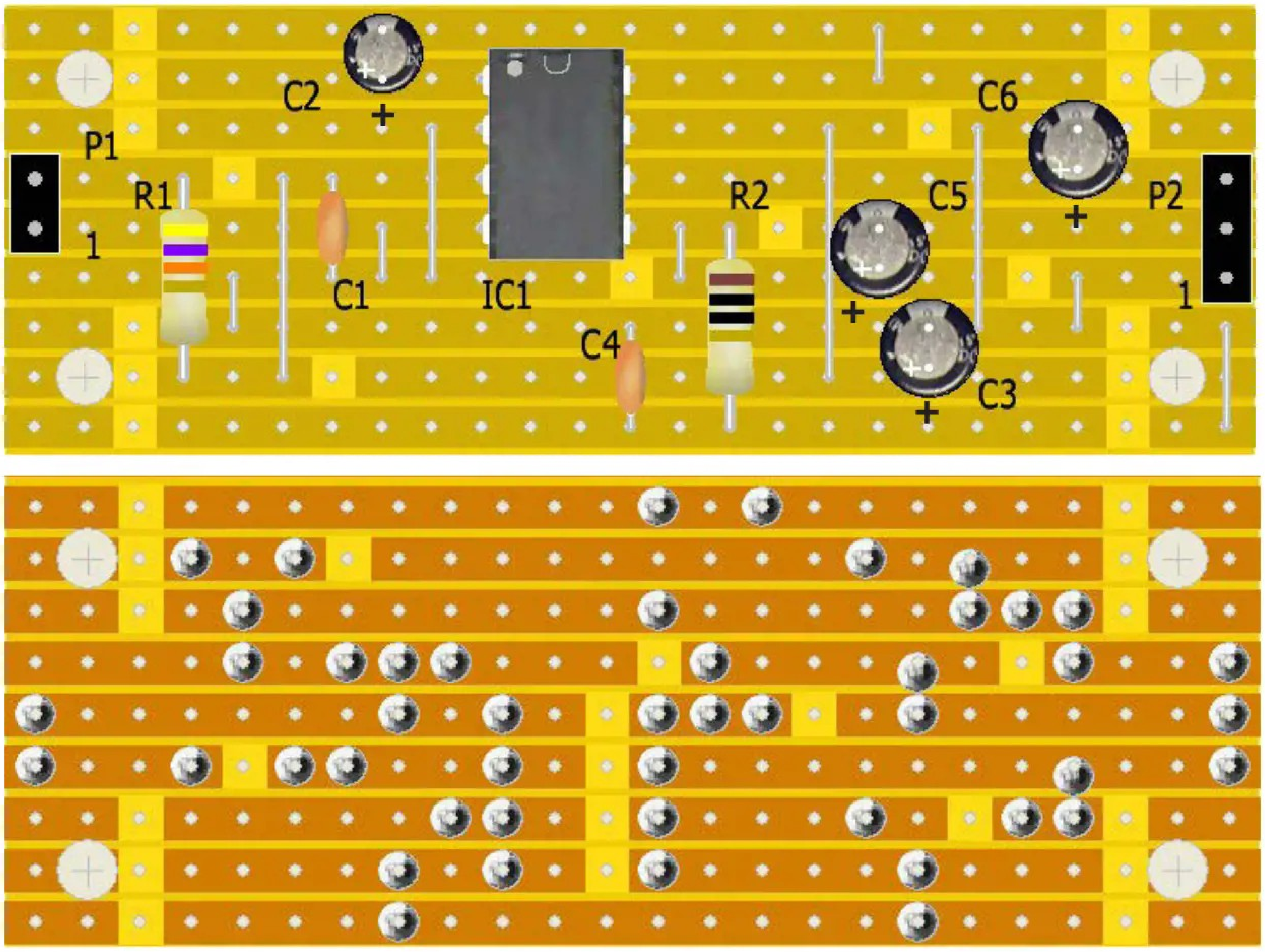
The 9V supply is necessary for the single-chip audio amplifier responsible for powering the loudspeaker. The receiver incorporates the same hand-wound coil and ferrite rod antenna that were used with last month's practical project.

The portable AM receiver uses two perforated copper stripboard modules, each measuring 25 × 64mm and arranged in nine strips each of 24 rows. One board is reserved for the RF circuitry (Fig.2.22), while the other is for the AF amplifier (Fig.2.23).

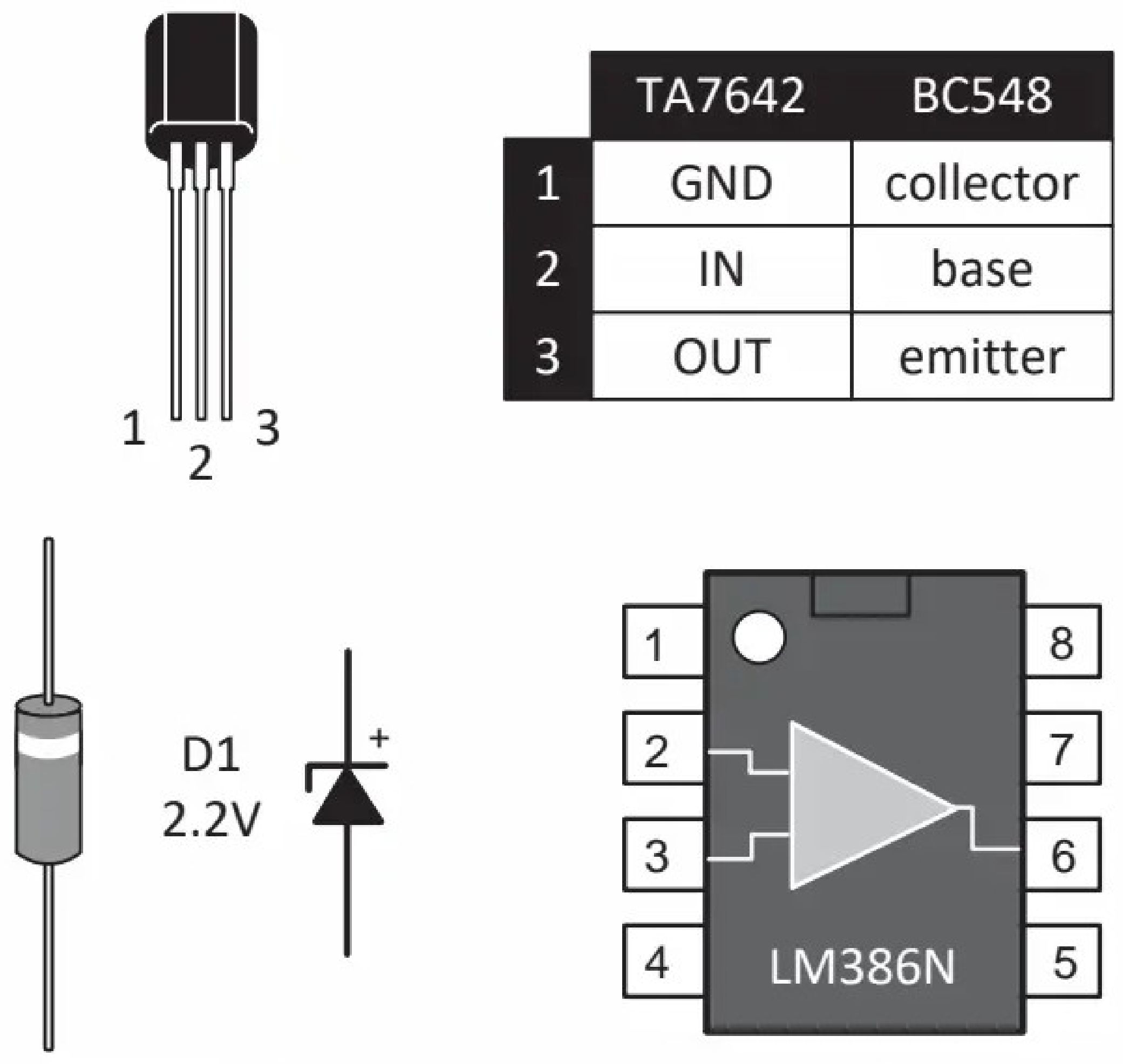
As with last month's project, tuning is managed by L1 and VC1, with L1 also serving as the antenna. L1 comprises 54 turns of 0.55mm ECW close-wound over a 70mm length of heatshrink sleeving, into which the ferrite rod is slid. The winding ends are secured with PVC tape.

The resulting winding will have an inductance of around 200µH, and with the 270pF variable capacitor, will provide a tuning range extending from around 700kHz to 1.6MHz (see last month's Hands-On feature for details). The precise value for VC1 is not critical, but a maximum value of 270-350pF will prove ideal.

IC1 is a TA7642 radio receiver chip housed in a TO-92 package



Figs.2.26 & 2.27: the component layout (top) and track view (below) for the AF module.



	TA7642	BC548
1	GND	collector
2	IN	base
3	OUT	emitter

Fig.2.28: the pin connections for the semiconductor devices used in our radio.

(see last month's article for details). The supply to the TA7642 is regulated at around 1.5V by TR1 and D1. The AF module is based on a universally available LM386N 8-pin audio amplifier, IC2, in a DIL package. This chip produces reasonable audio quality at ample volume for domestic and portable listening.

The component layouts for the two modules (viewed from the top) are shown in Figs.2.24 & 2.26, while the corresponding track layouts (viewed from below) are in Figs.2.25 & 2.27. The required track breaks can be made using a spot face cutter or small drill bit, and the links on the upper side of the boards are made using short lengths of tinned copper wire.

The pin connections for the semiconductor devices are shown in Fig.2.28.

When completed, the two stripboard modules, tuning components (L1 and VC1), volume control, RF gain control, loudspeaker, and battery holders can be mounted into an ABS enclosure of your choice, as shown in Fig.2.29. This can be the same enclosure used for last month's Hands-On project.

Connections to the off-board components (L1, VC1, VR1, VR2, etc) are made using the four 0.1-inch (2.54mm) pitch headers (shown as P1 and P2 on each board), together with short lengths of hook-up wire.

Testing

As always, it's important to check the stripboard and internal wiring before applying power. When these checks are complete, insert six AA cells into the holders. Switch on and advance the volume

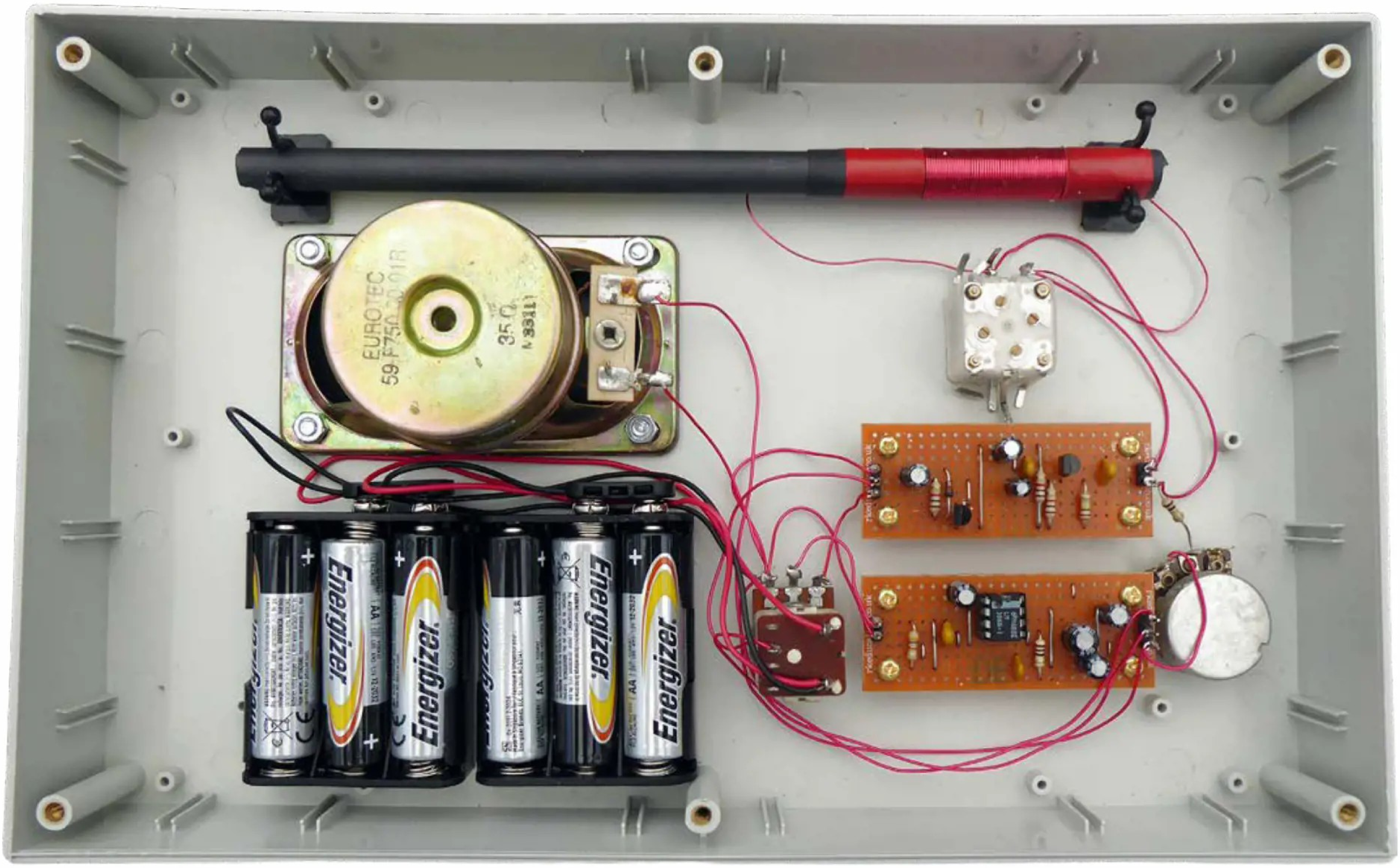


Fig.2.29: the internal assembly and wiring detail of the radio.

control. You should immediately hear some noise from the loudspeaker, and a quick sweep of VC1 should reveal several strong broadcast signals.

If that's not the case, switch off and recheck each board and wiring. For comparison, and to assist fault-finding, Table 2.2 provides the test voltages obtained from our prototype.

Using it

As with last month's Hands-On project, our portable AM receiver is extremely sensitive and can be susceptible to local RF noise sources like computers, TVs and many other electronic devices. For best results, you will need to keep the ferrite antenna well away from other devices.

The ferrite rod is directional; the strongest signal will be obtained when the antenna is side-on to the bearing of the station being received.

These directional properties can also be advantageous, allowing you to null or minimise the effects of local noise and interference. Simply rotate the antenna until you find the position with the least interference and best signal!

During the day, you will normally be able to receive several medium-wave broadcast signals at good strength.

Noticeably more signals will be present during the hours of darkness. In the UK, BBC Radio 5 Live broadcasts on 693kHz and 909kHz nationally, and should be received at good strength, as will Lyca Radio (broadcasting on several MW frequencies from both London and Manchester).

Live coverage of many sporting events as well as news and discussions is available from TalkSPORT on 1053kHz, 1089kHz, 1071kHz and 1107kHz. With the receiver placed in a favourable position, all these stations should be received easily and at good loudspeaker volume.

Coming up!

Part 3 of our Teach-In series will introduce signals and modulation while delving into basic arrangements for CW, AM and FM transmitters.

We will also explore RF semiconductor devices and show how they can be used to construct fixed and variable frequency oscillators in our two Hands-On projects.

Join us then! PE

Parts List – Improved MW Radio Receiver

RF Module
1 25 × 64mm piece of stripboard (9 × 24 holes)
1 3-pin male 0.1in/2.54mm header (P1)
1 2-pin male 0.1in/2.54mm header (P2)

Semiconductors
1 TA7642 single-chip radio receiver (IC1)
1 BC548 small-signal NPN transistor (TR1)
1 2.2V 300mW zener diode (D1)

Capacitors
2 10µF 16V radial electrolytic (C4, C5)
1 2.2µF 25V radial electrolytic (C3)
1 100nF 50V ceramic (C2)
1 10nF 50V ceramic (C1)

Resistors (all ¼W axial, 5% or better)
1 100kΩ (R1)
3 1kΩ (R2-R4)

AF Module
1 25 × 64mm piece of stripboard (9 × 24 holes)
1 3-pin male 0.1in/2.54mm header (P1)
1 2-pin male 0.1in/2.54mm header (P2)

Semiconductors
1 LM386N 8-pin DIL audio amplifier (IC1)

Capacitors
1 220µF 16V radial electrolytic (C5)
3 10µF 16V radial electrolytic (C2, C3, C6)
1 100nF 50V ceramic (C4)
1 1nF 50V ceramic (C1)

Resistors (all ¼W axial, 5% or better)
1 47kΩ (R1)
1 10Ω (R2)

Off-board components
1 ABS enclosure (optional; see text)
1 270-350pF miniature solid-dielectric variable capacitor (VC1)
1 200mm-long, 10mm diameter ferrite rod (L1)
1 35Ω miniature loudspeaker (LS1)
1 250kΩ linear carbon track potentiometer (VR1)
1 10kΩ logarithmic carbon track potentiometer (VR2)
2 3×AA cell battery holders
2 battery clip connectors
8 brass or nylon M3 × 10mm hex stand-off spacers
8 short M3 panhead machine screws

Table 2.2 – Expected voltages

Device	Pin	Voltage
IC1	1	0V
	2	1.3V
	3	1.1V
TR1	C	9.4V
	B	2.2V
	E	1.6V
IC2	1	1.3V
	2	0V
	3	0V
	4	0V
	5	4.8V
	6	9.4V
	7	4.8V
	8	1.3V

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Constructional Project

Variable Speed Drive Mk2

For Induction Motors

Part 1 by Andrew Levido

This new VSD significantly improves on our previous design. It's more compact, lighter, better cooled, and more efficient. It has better safety margins (making it more robust) and some new features.

Silicon Chip magazine published my induction motor speed controller more than ten years ago (April & May 2012; siliconchip.au/Series/25). That was a major project that was very popular but it had some drawbacks, including marginal cooling. So I thought it was time for an improved version.

Some critical components used in the old design, notably the integrated IGBT/driver module, are now obsolete. We took advantage of some other technological advances to make this unit smaller, more efficient and easier to build.

We have used a compact tunnel heatsink, with active cooling via a small DC fan, to significantly decrease the size and weight of the unit compared to its predecessor. Other things contribute to its compactness, like using switch-mode AC/DC converters rather than transformers and active discharge of the HV capacitor bank, rather than bulky power resistors acting as bleeders.

Functionally, the speed controller is similar to the previous unit in that it is designed to run single-phase shaded pole or permanent split capacitor (PSC) motors rated up to 1.5kW (2 horsepower) or any three-phase induction motor of a similar rating, as long as it can be configured for 230V operation (most can).

Like all such drives, the Variable Speed Drive (VSD) described here is not generally suitable for use with induction motors with centrifugal switches, since the start windings in these motors are not rated for continuous operation. You can read more about that in the panel on page 23 about how induction motors work.

An induction motor's power rating describes the output power at the shaft, not the electrical input required to run it. For example, your average 1.5kW single-phase induction motor draws a full-load current of 8.9A with a power factor of 0.95. That means the real power input is around 2kW, suggesting an efficiency of about 73%.

For a three-phase 1.5kW motor, the typical full-load phase current will be about 5.5A at a power factor of 0.82, giving a similar real power input.

This VSD can deliver around 9A continuously in single-phase mode and around 5.5A per phase in three-phase mode. It can deliver twice this current on a very short-term basis when starting the motor. The output frequency can be varied from 0.5Hz to 50Hz with a set point resolution of 0.25Hz. When ramping between set point frequencies, the frequency steps are even smaller than this.

Features

The basic topology of the power electronics is shown in the upper part of Fig.1. Power from the mains is rectified and filtered to create a DC 'bus' voltage of around 330V. This DC bus voltage is then pulse-width modulated by an IGBT bridge to produce the desired output voltage.

Only two of the output IGBT pairs (the U and V phases) are used in single-phase mode. This part of the circuit operates at mains voltages and stores considerable energy. Contact with any part of this can be lethal – so exercise care.

Because of this risk, the user controls, shown diagrammatically in the lower part of Fig.1, are isolated from the power circuit and near Earth potential. The board has six DIP switches for setting the operation modes. These are only read at start-up, so changing any of these while the speed controller is powered up has no effect until the next restart.

If the first of these switches is closed, it selects three-phase motor operation; otherwise, the controller operates in single-phase mode.

The second and third switches are related to pool pump operation. Running a pool pump at a lower speed can

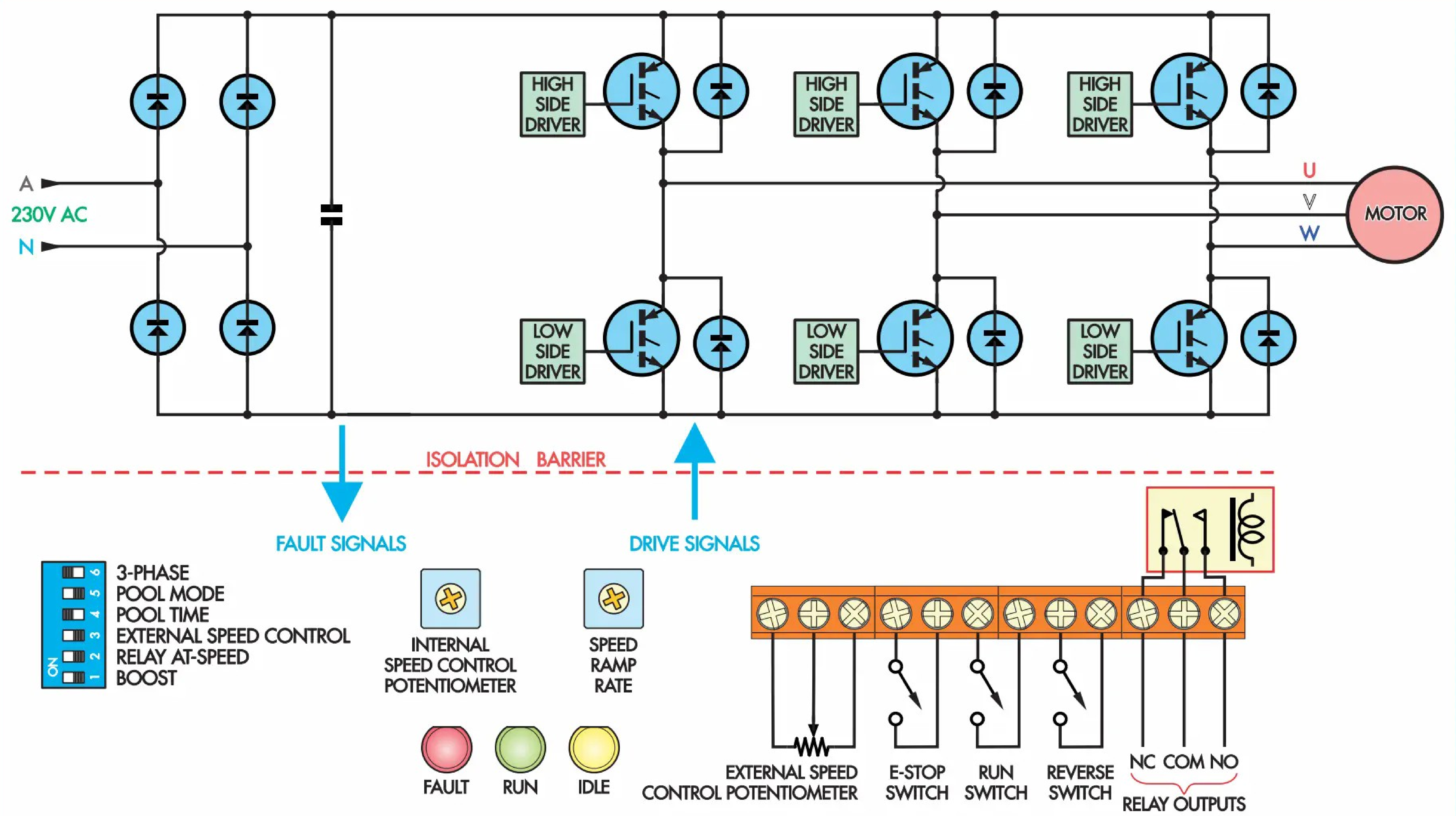

A
230V AC
N
HIGH SIDE DRIVER
HIGH SIDE DRIVER
HIGH SIDE DRIVER
LOW SIDE DRIVER
LOW SIDE DRIVER
LOW SIDE DRIVER
U
V
W
MOTOR
ISOLATION BARRIER
FAULT SIGNALS
DRIVE SIGNALS
ON
1 2 3 4 5 6
3-PHASE
POOL MODE
POOL TIME
EXTERNAL SPEED CONTROL
RELAY AT-SPEED
BOOST
INTERNAL SPEED CONTROL POTENTIOMETER
SPEED RAMP RATE
FAULT
RUN
IDLE
EXTERNAL SPEED CONTROL POTENTIOMETER
E-STOP SWITCH
RUN SWITCH
REVERSE SWITCH
NC COM NO
RELAY OUTPUTS

Fig.1: a basic overview of how the VSD works. At this 'zoomed out' level, it's similar to the previous IMSC (Induction Motor Speed Controller); the mains is rectified and charges a DC capacitor bank. The voltage from that bank is chopped by three IGBT half-bridges and applied to the motor windings, with the isolated control circuitry shown below.

save a considerable amount of energy in cases where the pump has to run for a long time – such as when a saltwater chlorinator is used. These typically must run for four to eight hours daily in summer to produce sufficient chlorine.

Under these circumstances, running the pump at 50% or even 75% of full speed can save a lot of energy. Just make sure the pump speed is high enough to keep the chlorinator cells covered and that the whole water volume is turned over at least once during each daily cycle.

If you operate a pool pump at reduced speed, it can be a good idea to run the pump for a short time at full speed first, to ensure the pump is primed and to purge any air from the system. That is the purpose of the pool pump mode.

If the Pool Mode DIP switch is closed, the VSD will initially ramp the motor up to full speed and hold it there briefly before ramping to whatever operating speed the user has set. The Pool Time DIP switch controls the duration of this full-speed period. If left open, the full-speed period is about 30 seconds; if it is closed, the time is extended to five minutes.

The Pool Mode and Pool Time switches are ignored if three-phase mode is selected.

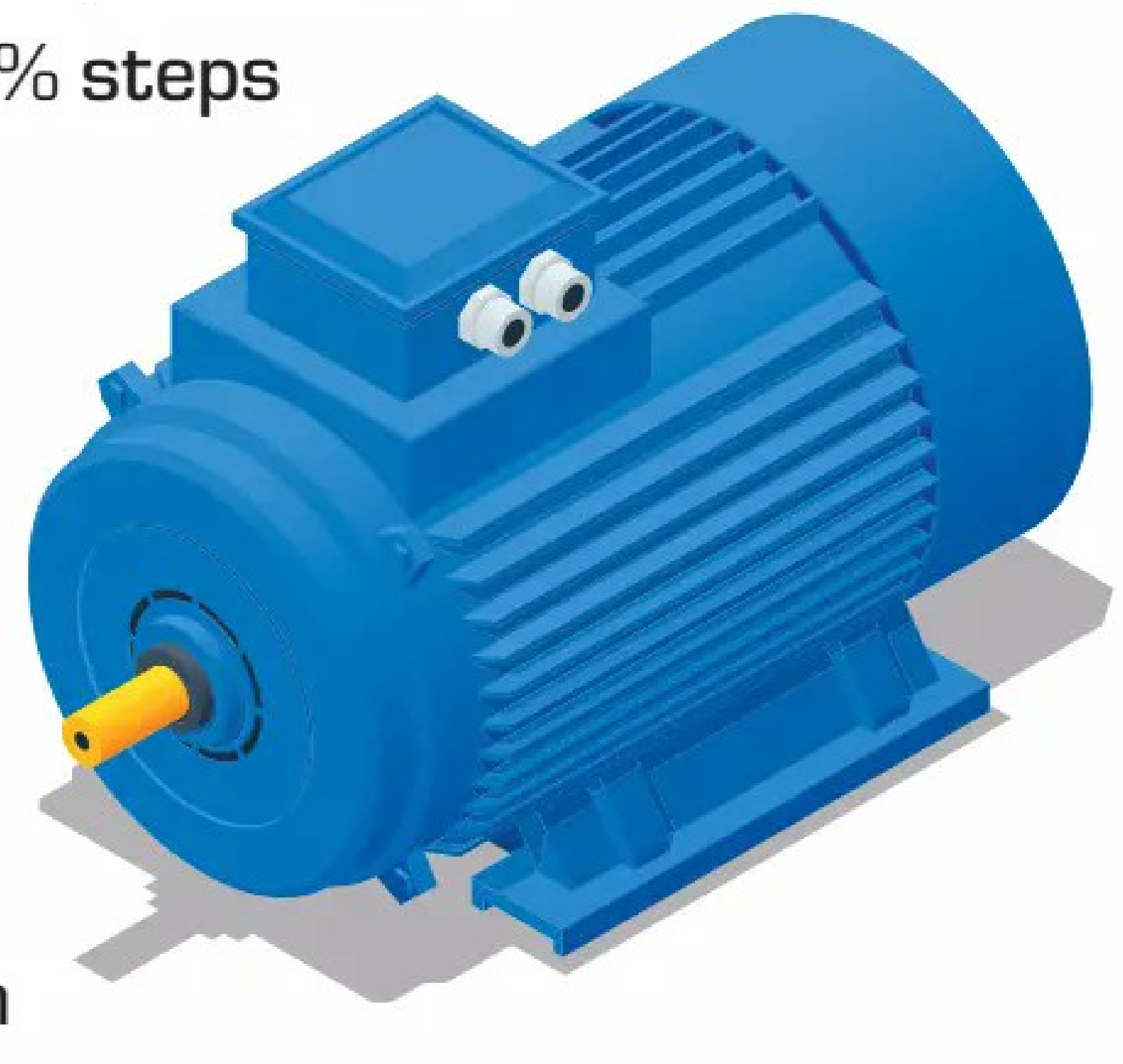
The speed control signal can come from either an onboard trimpot or an external potentiometer/control voltage. The latter option is selected by closing the External Speed DIP switch. Alongside the speed input terminal, 5V reference and ground terminals are provided for use with an external pot.

The reference can comfortably source 10mA, so any pot with a resistance of 500Ω or more can be used. You can also feed a 0-5V signal into this terminal to control the motor speed from an external device. The common terminal for the speed control is referenced to the mains Earth.

In addition to the internal speed control pot, there is a second trimpot

Variable Speed Drive Features & Specifications

- » Can drive single-phase shaded pole or PSC motors up to 1.5kW
- » Can drive three-phase 230V induction motors up to 1.5kW
- » Speed range: 1% to 100% of full speed in 0.5% steps
- » Runs from a standard 10A GPO
- » Inbuilt mains EMI/RFI filter
- » Robust inrush current limiting
- » Higher efficiency than our previous design
- » Fan-based cooling for critical components
- » Uses standard, discrete IGBTs for switching
- » Compact and lightweight
- » Over-current and over-temperature shutdown
- » Pool pump mode
- » Three-phase motors can be reversed at any time (they will slow down, stop, reverse and speed back up)
- » Adjustable speed ramp rate
- » Internal or external controls for speed, on/off and emergency stop
- » Relay outputs that switch when the motor is up to speed or on a fault



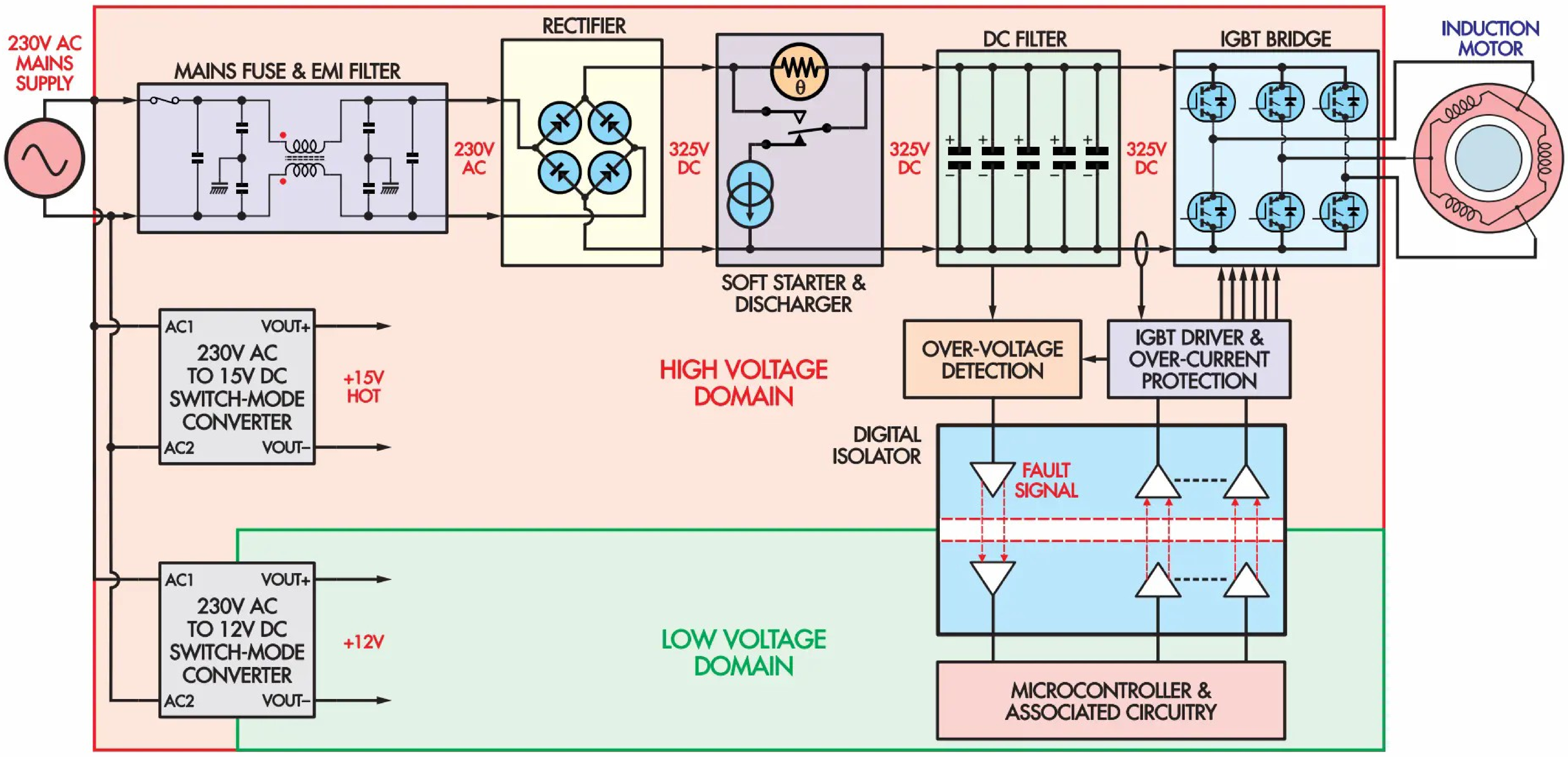


Fig.2: a somewhat more detailed view of how the VSD works. The soft starter & discharger block limits the inrush current into the capacitor bank when power is first applied and ensures that the bank discharges quickly when mains power is lost. Two similar AC-DC converters supply power to the 'hot' and isolated sections, with an eight-channel digital isolator bridging them.

to set the ramp rate. This controls how quickly the motor speed changes. The ramp rate can be set between three and 60 seconds for a ramp from zero to full speed. The longer ramp times may be necessary for high-inertia loads.

To get the speed controller to start, both the Run and E-Stop circuits must be closed or 12V fed into the relevant terminals from some external source. Opening the emergency stop (E-Stop) terminals immediately switches the IGBTs off, letting the motor freewheel to a stop. Opening the Run circuit causes the motor speed to ramp down to zero before the IGBTs are switched off.

The final external input is the Reverse control. This is only relevant in three-phase mode, and it sets the direction of rotation of the motor, effectively changing the phase sequence at the output. If you switch to the opposite direction while the motor is running, it will ramp down to zero, pause for two seconds, then ramp up again in the new direction.

Three LEDs indicate the VSD's operating status. The green LED indicates that the motor is running. It flashes quickly when the motor is ramping up or down and is illuminated steadily when the preset speed is reached. During the pool pump full-speed period, the green LED flashes slowly.

The yellow LED indicates that the speed controller is in idle mode. This means the IGBTs are off, but the VSD is ready to run once the E-Stop and Run switches are closed and a non-zero speed signal is applied.

The red LED indicates a fault condition. If just the red LED is illuminated, the fault is either an overcurrent trip or the DC bus voltage has risen too high. If the red and yellow LEDs are both lit, the heatsink temperature has become dangerously high. Either way, the fault can be reset by cycling power or toggling the E-Stop switch (opening then closing it) after the fault has cleared.

An output relay (RLY2) provides a set of uncommitted isolated changeover contacts that the user can employ as they see fit. The At-Speed DIP switch configures the relay function. If the DIP switch is open, the relay activates when a fault occurs. If closed, the relay activates when the motor has reached the preset speed.

The Boost DIP switch increases the motor voltage at very low speeds. You may need to switch this in to reliably start constant-torque loads such as displacement pumps, conveyers or hoists. Some pool pumps may also require this boost since the pump seals can sometimes become 'sticky' if the pump has been stationary for some time.

How it works

Fig.2 is a block diagram of the VSD showing the two distinct power domains. The high-voltage section containing the power electronics is shown in red, while the low-voltage part with the control circuitry is shown in green.

As we step through the full circuit (Fig.3), it may be helpful to refer to this diagram as well. The mains input first passes through a 10A slow blow fuse, F1 – a last line of defence in case of a catastrophic failure. It then passes through an EMI filter consisting of the common-mode inductor L1 and six capacitors.

The EMI filter is there to minimise the high-frequency artefacts (of which there are plenty in a circuit of this type) making their way back to the mains supply.

The mains supply is then rectified by a full-bridge rectifier, BR1, and applied to the five parallel DC bus capacitors via a soft start/discharge circuit. Thermistor NTC1, which has a resistance of around 10Ω when cold, limits the capacitor bank inrush current to about 35A peak.

We use a specialised inrush-limiting thermistor here because it would be difficult to guarantee the reliability of a generic power resistor in this application. The thermistor used here is rated for a maximum capacitor inrush energy of 150J. The maximum energy that our capacitor bank can store is 110J (from $E = ½CV^2$) if the mains voltage is at its upper limit of 260V.

Variable-Speed Drive Mk2, part one

SPEED
STOP
RUN
REVERSE
MORE MAGAZINES: XSAVA.XYZ

The third (black) cable gland is for wiring to an optional external controller, which can be as simple as the one shown here.

The thermistor's resistance drops dramatically as it heats up, and it can continuously pass 15A – more than enough for this application. However, unlike the original controller, we have chosen to short it out with relay RLY1, which closes once the capacitors are charged. This simultaneously disconnects the capacitor discharge section when the speed controller is operating.

Shorting out the NTC thermistor has a few advantages. Firstly, it increases efficiency and reduces heat dissipation in the case due to the thermistor's resistance. It also cools down more quickly after the unit is switched on, so it will effectively reduce the inrush current if the unit is switched off and then (almost) immediately on again.

The capacitor discharge circuit is also an upgrade from the previous design. There, we used three bulky 5W power resistors, which resulted in a discharge time of about 90 seconds and continuous power dissipation approaching 10W (a complete waste).

This time, we have used a constant-current discharge circuit based around transistors Q7 and Q8. This discharges the capacitors at a nominal 50mA, taking around 10 seconds, making the unit much safer to work on. Switching it out during operation again improves efficiency and greatly reduces the heating inside the case.

Adding RLY1 has eliminated a total of about 20W of continuous power dissipation compared to the previous design.

The capacitor bank itself deserves a few words. The input current of any circuit like this, which rectifies and filters the mains, is very 'spikey' as the rectifier diodes only conduct at the very peak of the mains. This results in a pretty terrible input power factor and very high levels of ripple current in the capacitors. The current flowing out of the capacitors to the motor also contributes.

A simulation (this is very hard to calculate any other way) showed this ripple to be around 10A RMS in total, or 2.0A RMS per capacitor. Therefore, it is essential to use capacitors designed for a 100Hz ripple current of at least 2A, like the Nichicon caps specified in the parts list.

After the filter capacitors, there is a 15mΩ current-sensing resistor (more on this later) and more EMI suppression via another set of three X2/Y2 capacitors. These help to shunt any high-frequency artefacts on the DC bus to ground or Earth.

Another big difference between this design and the previous Induction Motor Speed Controller (IMSC) is the use of discrete IGBTs (Q1 through Q6) and a separate driver chip (IC2) instead of an integrated power module.

The DGTD65T15H2TF IGBTs used here are rugged devices rated at 650V/30A and specifically designed for motor drive use. They include an anti-parallel diode with similar ratings, and come in an isolated TO-220 case.

The latter is important since we want to use an Earthed heatsink for safety and don't want to have to fuss with insulating washers and the like.

The diode bridge and discharge Mosfet, Q8, are also mounted on the heatsink; all use isolated packages for maximum convenience and safety.

Driving the IGBTs

The IGBTs are driven by a surprisingly inexpensive, specialised IGBT driver chip, the Infineon 6EDL04I06PT (IC2). The block diagram of this chip is reproduced in Fig.4. For each of the three phases, there are two logic-level inputs, one for the high-side IGBT and one for the low-side. In addition, a global enable pin (EN) must be high for any of the drivers to be active.

These inputs pass through a noise filter to some logic that prevents both high-side and low-side IGBTs in the same phase from being switched on at once. The logic also ensures there is a short dead time when switching between high-side and low-side transistors or vice versa. About 310ns in length, this is sufficient to give one

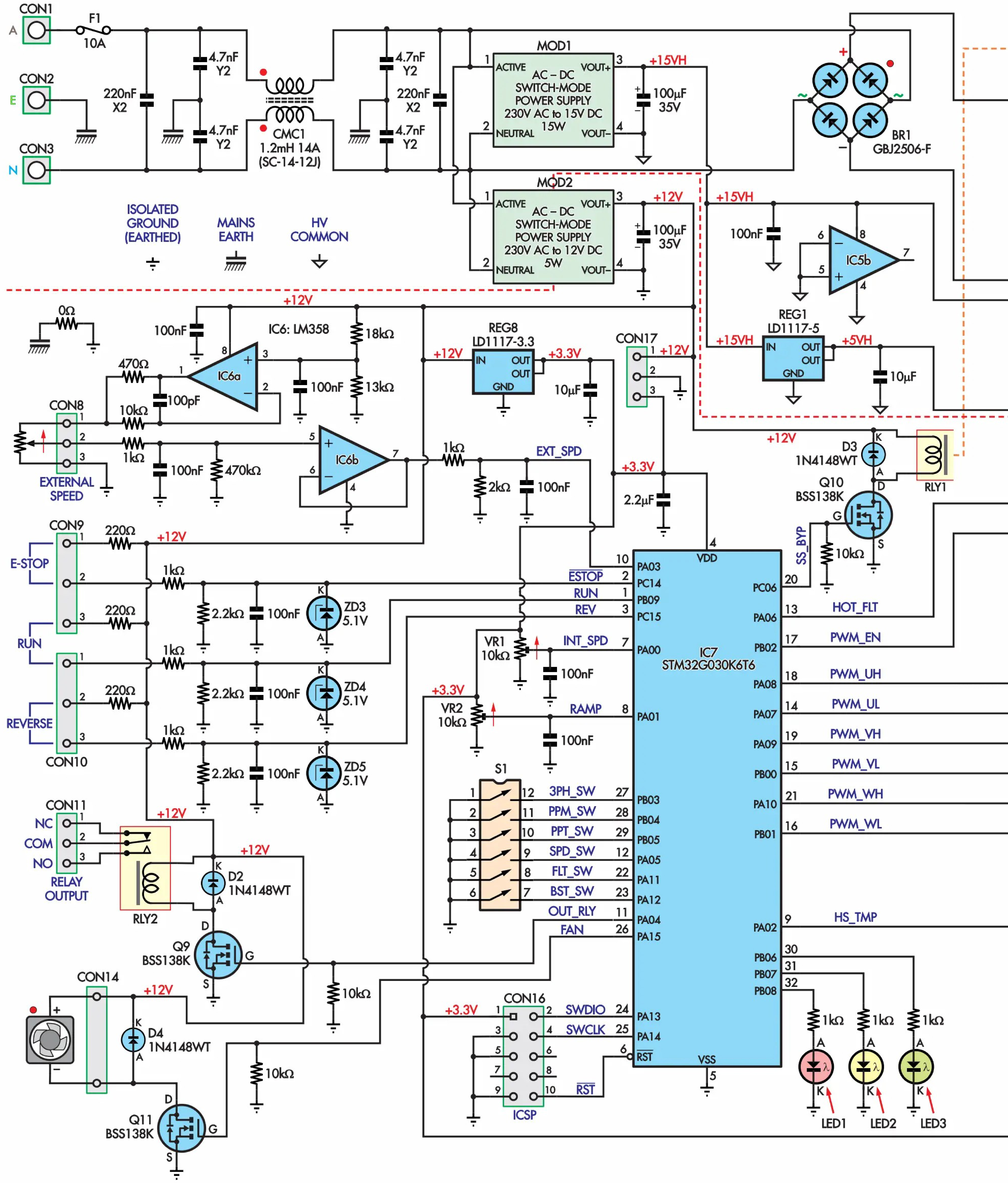
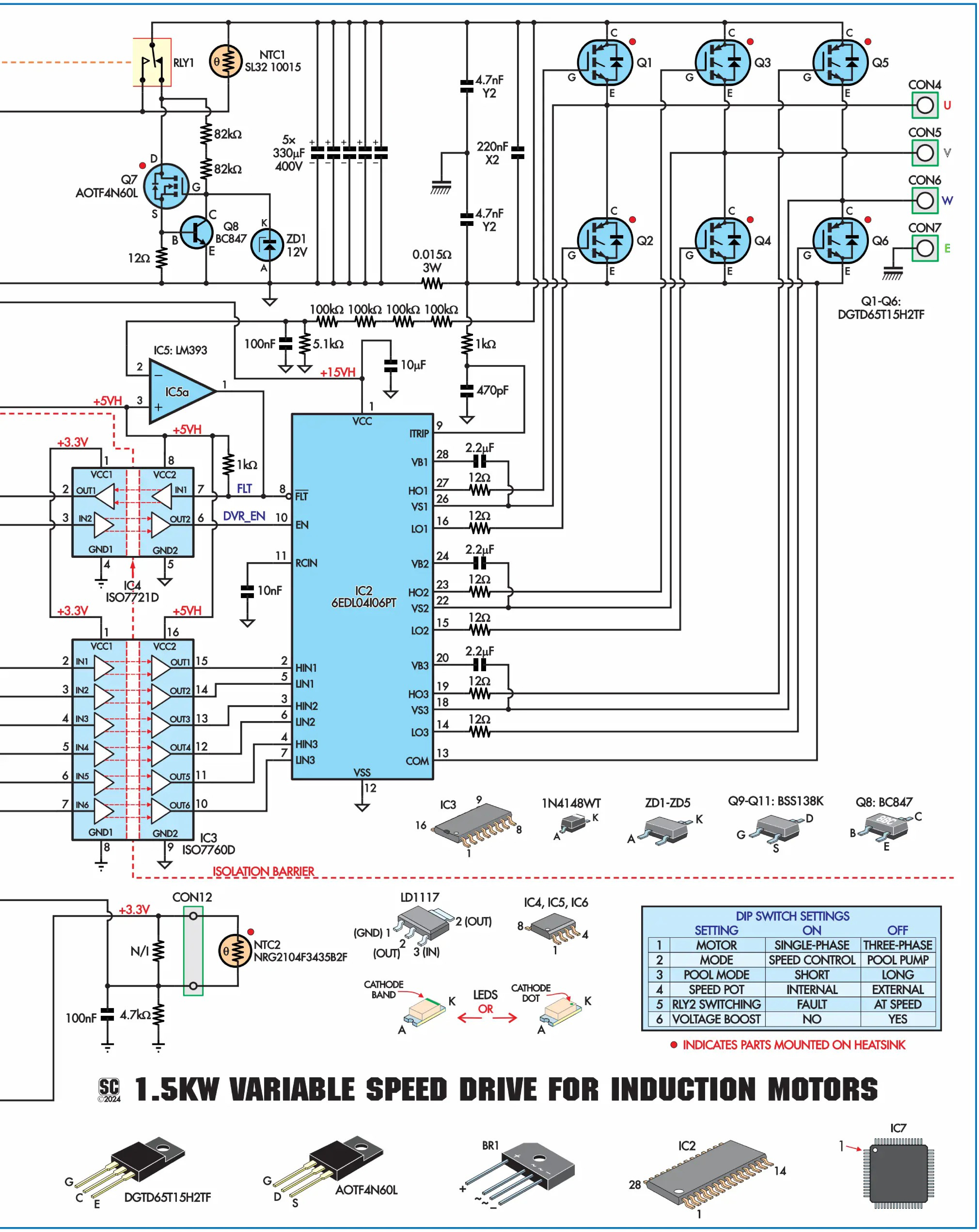

CON1
A
F1
10A
CON2
E
CON3
N
220nF
X2
4.7nF
Y2
4.7nF
Y2
CMC1
1.2mH 14A
(SC-14-12J)
4.7nF
Y2
220nF
X2
4.7nF
Y2
MOD1
ACTIVE
AC – DC
SWITCH-MODE
POWER SUPPLY
230V AC to 15V DC
15W
NEUTRAL
VOUT+
VOUT–
+15VH
100µF
35V
MOD2
AC – DC
SWITCH-MODE
POWER SUPPLY
230V AC to 12V DC
5W
+12V
100µF
35V
BR1
GBJ2506-F
+15VH
100nF
IC5b
REG1
LD1117-5
IN
OUT
GND
+5VH
10µF
ISOLATED
GROUND
(EARTHED)
MAINS
EARTH
HV
COMMON
0Ω
100nF
IC6: LM358
18kΩ
13kΩ
470Ω
100pF
IC6a
100nF
10kΩ
1kΩ
CON8
EXTERNAL
SPEED
100nF
470kΩ
IC6b
1kΩ
2kΩ
100nF
EXT_SPD
REG8
LD1117-3.3
+3.3V
10µF
CON17
+3.3V
2.2µF
+12V
D3
1N4148WT
Q10
BSS138K
RLY1
SS_BYP
10kΩ
CON9
E-STOP
RUN
REVERSE
CON10
220Ω
1kΩ
2.2kΩ
100nF
ZD3
5.1V
ZD4
5.1V
ZD5
5.1V
VR1
10kΩ
VR2
10kΩ
ESTOP
RUN
REV
INT_SPD
RAMP
IC7
STM32G030K6T6
VDD
VSS
PA03
PC14
PB09
PC15
PA00
PA01
PB03
PB04
PB05
PA05
PA11
PA12
PA04
PA15
PA13
PA14
RST
PC06
PA06
PB02
PA08
PA07
PA09
PB00
PA10
PB01
PA02
PB06
PB07
PB08
HOT_FLT
PWM_EN
PWM_UH
PWM_UL
PWM_VH
PWM_VL
PWM_WH
PWM_WL
HS_TMP
S1
3PH_SW
PPM_SW
PPT_SW
SPD_SW
FLT_SW
BST_SW
OUT_RLY
FAN
CON11
NC
COM
NO
RELAY
OUTPUT
RLY2
D2
1N4148WT
Q9
BSS138K
10kΩ
CON14
D4
1N4148WT
10kΩ
Q11
BSS138K
CON16
SWDIO
SWCLK
RST
ICSP
1kΩ
LED1
LED2
LED3

Fig.3: the complete VSD circuit. The red dashed line is the isolation barrier; note how RLY1 also bridges it. Comparator IC5a's output goes low if the capacitor bank voltage gets too high, while IC2 pulls the same FLT line low if an overcurrent condition is detected. Either way, the drive to the IGBTs shuts down.



RLY1
NTC1
SL32 10015
82kΩ
82kΩ
5× 330μF 400V
4.7nF Y2
220nF X2
4.7nF Y2
Q7
AOTF4N60L
D
G
S
Q8
BC847
B
C
E
12Ω
ZD1
12V
K
A
0.015Ω 3W
Q1
Q2
Q3
Q4
Q5
Q6
C
G
E
CON4
U
CON5
V
CON6
W
CON7
E
Q1-Q6:
DGTD65T15H2TF
100kΩ 100kΩ 100kΩ 100kΩ
100nF
5.1kΩ
1kΩ
10μF
470pF
+15VH
IC5: LM393
IC5a
+5VH
+3.3V
1kΩ
FLT
DVR_EN
VCC1
VCC2
OUT1
IN1
IN2
OUT2
GND1
GND2
IC4
ISO7721D
10nF
VCC
ITRIP
VB1
HO1
VS1
LO1
VB2
HO2
VS2
LO2
VB3
HO3
VS3
LO3
COM
VSS
EN
RCIN
HIN1
LIN1
HIN2
LIN2
HIN3
LIN3
IC2
6EDL04I06PT
2.2μF
12Ω
IN3
IN4
IN5
IN6
OUT3
OUT4
OUT5
OUT6
IC3
ISO7760D
ISOLATION BARRIER
IC3
1N4148WT
ZD1-ZD5
Q9-Q11: BSS138K
Q8: BC847
CON12
N/I
NTC2
NRG2104F3435B2F
100nF
4.7kΩ
LD1117
(GND) 1
2 (OUT)
(OUT)
3 (IN)
IC4, IC5, IC6
CATHODE BAND
CATHODE DOT
LEDS
OR
K
A
DGTD65T15H2TF
AOTF4N60L
BR1
IC2
IC7
SC ©2024

DIP SWITCH SETTINGS			
	SETTING	ON	OFF
1	MOTOR	SINGLE-PHASE	THREE-PHASE
2	MODE	SPEED CONTROL	POOL PUMP
3	POOL MODE	SHORT	LONG
4	SPEED POT	INTERNAL	EXTERNAL
5	RLY2 SWITCHING	FAULT	AT SPEED
6	VOLTAGE BOOST	NO	YES

● INDICATES PARTS MOUNTED ON HEATSINK

1.5KW VARIABLE SPEED DRIVE FOR INDUCTION MOTORS

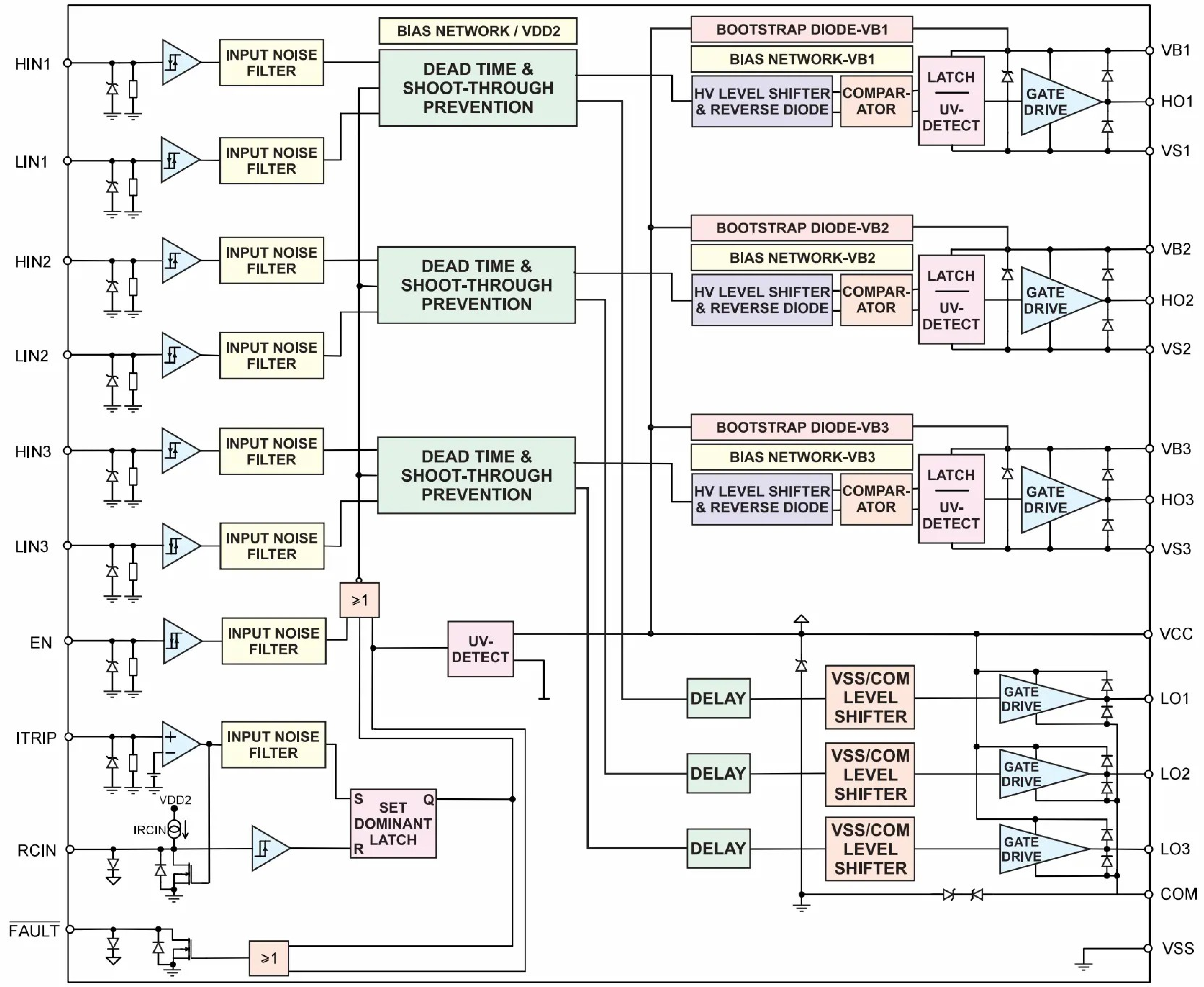

HIN1
LIN1
HIN2
LIN2
HIN3
LIN3
EN
ITRIP
RCIN
FAULT
INPUT NOISE FILTER
BIAS NETWORK / VDD2
DEAD TIME & SHOOT-THROUGH PREVENTION
BOOTSTRAP DIODE-VB1
BIAS NETWORK-VB1
HV LEVEL SHIFTER & REVERSE DIODE
COMPAR-ATOR
LATCH
UV-DETECT
GATE DRIVE
BOOTSTRAP DIODE-VB2
BIAS NETWORK-VB2
BOOTSTRAP DIODE-VB3
BIAS NETWORK-VB3
VB1
HO1
VS1
VB2
HO2
VS2
VB3
HO3
VS3
VCC
LO1
LO2
LO3
COM
VSS
≥1
UV-DETECT
DELAY
VSS/COM LEVEL SHIFTER
VDD2
IRCIN
S
R
Q
SET DOMINANT LATCH

Fig.4: a colourised and cleaned-up version of the internal block diagram from the 6EDL04I06PTXUMA1 IGBT driver data sheet. It provides all the functions we need to drive the six IGBTs and monitor the current draw in one package.

IGBT time to turn off before its opposite number begins to turn on.

The microcontroller also inserts dead time into the PWM signals, so this circuit provides some useful 'belts and braces' backup should something unexpected happen.

From there, the high-side signals go to three high-side IGBT gate drivers via level shifters. This is necessary because these gate drivers are referenced to the high-side IGBT's emitters via the VS1, VS2 and VS3 pins. In operation, these pins are switching alternatively between the negative side of the DC bus (when the low side IGBT is on) and the positive side of the bus (when the high-side IGBT is on).

Most of the circuitry in the high-voltage domain, including the IGBT driver's VSS pin, is referenced to the negative side of the DC bus. The circuit diagram shows this with a triangular 'ground' symbol. Do not confuse this with the common in the low voltage domain (shown with the usual ground symbol having three horizontal lines), which is referenced to mains Earth.

You will also notice a 'chassis Earth' symbol in a few places. This symbol refers specifically to mains Earth connections. It consists of two thick horizontal bars with a series of diagonal lines coming off the lower one.

Returning to IC2, the low-side drive signals are routed to the three low-side gate drivers via a delay block, which is necessary to match the delay introduced by the high-side level shifters. The low-side gate drivers are referenced to the COM pin, which is connected to the low-side IGBT emitters.

This COM signal can float a few volts up or down with reference to VSS (HV_COM) since there may be some voltage drop across the 15mΩ current shunt resistor and the PCB traces.

The IGBT driver is powered by a +15V supply applied to the V_{CC} pin. This supply is used for the logic and low-side drivers directly, but powers each high-side driver via three bootstrap circuits. These consist of internal bootstrap diodes connected between V_{CC} and three 2.2μF external capacitors connected to the VB1, VB2 and VB3 pins.

When a low-side IGBT is on, the corresponding high-side driver's bootstrap capacitor charges via its bootstrap diode. When the low-side driver is off, the diode is reverse biased, and the capacitor provides a floating power source for the high-side driver. An undervoltage lockout prevents the high-side driver from operating if its bootstrap voltage is not sufficient.

Overcurrent and overvoltage protection

The 6EDL04I06PT driver includes a trip circuit to protect the IGBTs from overloads or short circuits. This works by monitoring the voltage at the ITRIP pin and shutting down the drive to all

Single-Phase Induction Motors

With a 3-phase supply, achieving a rotating magnetic field is easily achieved by spacing the three windings around the rotor. Swap any two of the phases and the field will rotate in the opposite direction.

With a single-phase supply, the sole winding can only produce a pulsating field. There is no torque on the rotor when it is stationary, so it cannot start without some impulse to get it going. Once moving, the torque builds up. The motor will rotate equally well in either direction, depending on the sense of this initial kick.

There are a few different schemes to give this initial kick-start. Manufacturers have not adopted a common set of terms to describe their various approaches, so the whole topic is potentially confusing.

Below, we have summarised a few of the more common starting mechanisms:

Shaded Pole ✓

A shorted turn on the corner of the stator poles distorts the magnetic field to create a weak starting torque. Shaded pole motors are inefficient due to the shorted turn and so usually limited to low power motors such as found in small domestic fans and blowers. These motors can be used with a speed controller such as the one described here but generally that would be an expensive solution for a low-power device.

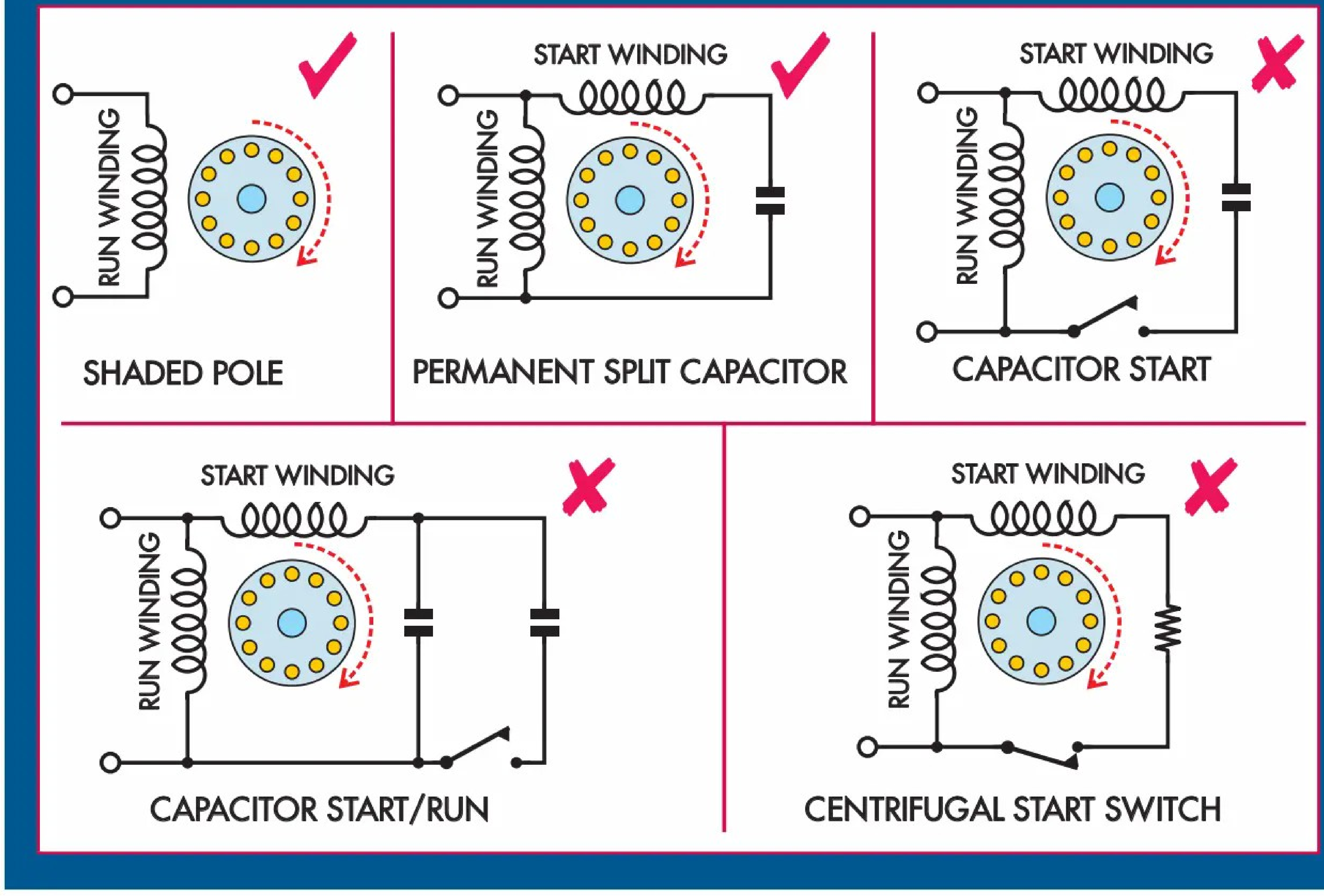
Permanent Split Capacitor ✓

A start winding in series with a capacitor produces a second, weaker field slightly out of phase with the main field. It is designed to draw a relatively modest current and rated for continuous operation.

Permanent Split Capacitor (PSC) motors have low starting torque and are very reliable since there is no centrifugal switch. Typically used for fans and centrifugal (pool & spa) pumps up to about 2kW, these are suitable for use with a speed controller.

Capacitor Start ✗

These are similar to the PSC motor in that a capacitor and start winding create a phase-shifted field for starting. The capacitor is larger and the start winding designed to draw significantly more current and therefore provides a much higher starting torque.



The start winding and capacitor are not rated for continuous operation and waste a lot of energy so are switched out by a centrifugal switch, typically at about 70% of full speed.

They are used for conveyors, large fans, pumps and geared applications requiring high starting torque. Capacitor Start motors are not suitable for speed control use because at lower speeds the centrifugal switch will close and the start winding or capacitor may burn out.

Capacitor Start/Run ✗

These are the “big guns” of single-phase motors and are used for machine tools, compressors, brick saws, cement mixers etc. They have a large start capacitor that is switched out by a centrifugal switch and a smaller run capacitor that is permanently connected to the start winding. They have very high starting torque and good overload performance.

For the same reason as the capacitor start motors, they cannot be used with variable speed drives. A 3-phase motor is recommended in these applications if speed control is desirable.

Centrifugal Start Switch ✗

Commonly used on small bench grinders and column drills, these motors arrange a phase-shifted field with a resistive winding. Again, the start winding is only rated for intermittent operation (due to its high resistance) and will burn out if operated continuously.

NOTE: in spite of the above warnings, some readers may want to try using the VSD with motors using a centrifugal switch to energise the start winding. The danger is that the start winding may be burnt out if it is energised for too long when operating at low speeds. There is also a risk that the over-current protection in the VSD will prevent normal operation.

WARNING: DANGEROUS VOLTAGES

This circuit is directly connected to the 230V AC mains. As such, most of the parts and wiring operate at mains potential. Contact with any part of these non-isolated circuit sections could prove fatal.

Note also that the circuit can remain potentially lethal even after the 230V AC mains supply has been disconnected! To ensure safety, this circuit MUST NOT be operated unless it is fully enclosed in a plastic case. Do not connect this device to the mains with the lid of the case removed. Do not touch any part of the circuit for at least 30 second after unplugging the power cord from the mains socket.

This is not a project for the inexperienced. Do not attempt to build it unless you understand what you are doing and are experienced working with high-voltage circuits.

IGBTs if the voltage exceeds 0.45V. We use this to monitor the voltage across the 15mΩ shunt resistor, giving a nominal trip current of 30A.

An RC low-pass filter consisting of a 1kΩ resistor and 470pF capacitor provides some immunity from false triggering due to noise.

If an overcurrent condition is detected, the gate drivers are switched off, and a fault signal is asserted on the open-drain $\overline{\text{FLT}}$ pin, pulling the FLT line low. After a short time, dictated by the value of the 10nF capacitor at pin 11, the gate drivers are re-enabled, and the fault output is de-asserted.

The 10nF value sets this time to about 20ms, long enough for the microprocessor to detect the fault condition, disable the IGBT driver and latch the fault state.

In addition to the overcurrent detection provided by the IGBT driver, there is also an external overvoltage detection circuit on the DC bus. This voltage can increase when a motor is decelerated due to regeneration. The voltage rise can become significant if the load has a lot of inertia. In the worst case, it could exceed the capacitors' voltage ratings.

A voltage divider consisting of four series 100kΩ resistors and a 5.1kΩ resistor to HV common reduces the bus voltage by a factor of about 80. If the divider's output reaches 5V, corresponding to a bus voltage of 400V, comparator IC5a's open-collector output will pull the FLT line low.

The overvoltage and overcurrent faults are therefore wire-ORed together to create a single fault signal that deactivates the IGBT drive of IC2 and is also transmitted across the isolation barrier (via IC4) to the microcontroller.

Parts List – Variable Speed Drive

1 double-sided PCB coded 11111241, 150 × 205mm, black solder mask
1 Hammond/Multicomp HM1112/RP1455 220 × 165 × 60mm enclosure **[Farnell 1635140]**
1 Zettler ZP05S1500WB mains to 15V DC 5W AC/DC converter (MOD1)
1 Zettler ZP05S1200WB mains to 12V DC 5W AC/DC converter (MOD2)
2 M205 PCB-mount fuse clips (for F1)
1 10A M205 slow-blow ceramic fuse (F1) **[Bel 5HT 10-R]**
1 vinyl M205 fuse cover/insulator (for F1) **[Keystone 3527C]**
1 SL32 10015 10Ω 15A NTC thermistor (NTC1)
1 NRG2104F3435B2F 10kΩ lug-mount NTC thermistor (NTC2)
1 1.2mH 14A toroidal common-mode choke (CMC1) **[Kemet SC-14-12J]**
2 J107F1CS1212VDC.45 12V DC coil 12A SPDT relays (RLY1, RLY2)
2 10kΩ mini top-adjust single-turn 3362P-style trimpots (VR1, VR2)
1 6-way DIP switch (S1) **[CUI DS01C-254-L-06BE]**
7 vertical PCB-mounting 5mm pitch 4.8mm male spade lugs (CON1-CON7)
4 3-way mini terminal blocks, 5.08mm pitch (CON8-CON11)
1 2×5-pin keyed shrouded SMD box header, 1.27mm lead pitch (CON16)
  [CNC Tech 3220-10-0300-00]
1 3-pin header, 2.54mm pitch (CON17)
1 100mm-long 40 × 40mm tunnel heatsink
  [AliExpress 1005006064507597 or AliExpress 1005006255161284]
1 40 × 40 × 20mm 12V DC 0.3m^3/minute maglev fan **[Sunon MF40201VX-1000U-A99]**
1 40mm fan guard & filter [Qualtek 09150-F/45]
1 10A mains extension cord
1 150mm length of 10A green/yellow striped wire
2 cable glands to suit the mains extension cord
7 4.8mm female spade crimp lugs to suit 1mm^2 wire
2 4.8mm female piggyback spade crimp lugs to suit 1mm^2 wire
1 100mm length of 8mm diameter blue heatshrink tubing
1 100mm length of 8mm diameter red heatshrink tubing
1 100mm length of 10mm diameter green/yellow striped heatshrink tubing
1 small cable gland (**optional**; for external control box)
1 external control box (**optional**; see separate parts list)
4 M3 × 25mm panhead machine screws
15 M3 × 10mm panhead machine screws
11 M3 spring washers
4 No.4 × 6mm self-tapping screws
1 small tube of thermal paste
1 small tube of superglue
small zip-lock cable ties
extra cabling required for connection to a 3-phase motor

Semiconductors
1 6EDL04I06PTXUMA1 high-voltage three-phase H-bridge gate driver, SOIC-28 (IC2)
1 ISO7760DW six-channel unidirectional digital isolator, wide SOIC-16 (IC3)

Power supply and isolation

The high-voltage domain circuitry is powered by a small modular AC-to-DC switch-mode converter that supplies 15V (designated +15VH on the circuit diagram) at 5W from the mains. 5V linear regulator REG1 produces the +5VH rail for the fault logic and the digital isolators.

The previous IMSC used a relatively large and heavy mains power transformer instead of a switch-mode supply. While there is an argument for preferring the simplicity of a transformer, these switch-mode supplies are less expensive, considerably smaller, lighter, and more efficient and allow the unit to operate from a wide range of mains supply voltages.

This brings us to another improvement on the earlier controller, which used opto-couplers to transmit the control signals across the isolation barrier. This design uses modern low-cost digital isolators. They work by modulating the input signal, passing it capacitively across an insulating barrier and demodulating it on the other side to reconstruct the original signal.

The ones we used here have an isolation voltage of 5000V RMS (somewhat less of a 'reinforced' rating, but still plenty for mains work) and support data rates up to 100Mbps.

You can get these digital isolators in all sorts of configurations. We use one with six channels, all going in the same direction (IC3) for the PWM signals, and one with two channels (IC4), one going in each direction, for the enable (EN) and fault (FLT) signals.

The supply voltages on each side do not have to be the same. We have used 5V logic on the high-voltage side and 3.3V logic on the isolated (control) side.

Control circuitry

The STM32G030K6T6 microcontroller (IC7) is the heart of the control circuitry. This has a 32-bit ARM Cortex

1 ISO7721D two-channel bidirectional digital isolator, SOIC-8 (IC4)
1 LM393AD dual differential comparator, SOIC-8 (IC5)
1 LM358AD dual single-supply op amp, SOIC-8 (IC6)
1 STM32G030K6T6 32-bit ARM microcontroller with 32KiB flash, programmed with 1111124A.HEX, LQFP-32 (IC7)
1 LD1117S50 5V low-dropout linear regulator, SOT-223 (REG1)
1 LD1117S33 3.3V low-dropout linear regulator, SOT-223 (REG8)
6 DGTD65T15H2TF 650V 30A IGBTs, TO-220FP (Q1-Q6)
1 AOTF4N60L 600V 4A N-channel Mosfet, TO-220FP (Q7)
1 BC847C 45V 100mA NPN transistor, SOT-23 (Q8)
3 BSS138K 50V 220mA N-channel logic-level Mosfets, SOT-23 (Q9-Q11)
3 M2012/0805 size LEDs; red, yellow & green (LED1-LED3)
1 BZX84-C12 12V 250mV zener diode, SOT-23 (ZD1)
3 BZX84-C5V1 5.1V 250mV zener diodes, SOT-23 (ZD3-ZD5)
1 GBJ2506-F 600V 25A SIL bridge rectifier (BR1)
3 1N4148WT 75V 300mA switching diodes, SOD-523 (D2-D4)

Capacitors *(all SMD M2012/0805 size 50V X7R unless noted)*
5 330μF 400V 105°C snap-in electrolytic, 30mm diameter, 40mm tall **[Nichicon LGW2G331MELB40]**
2 100μF 35V 105°C SMD electrolytic, 6.3mm diameter **[Nichicon UCD1V101MCL6GS]**
3 10μF 25V
4 2.2μF 25V
3 220nF X2 capacitors, 15mm lead pitch, 7mm wide **[EPCOS/TDK B32922C3224K000]**
6 4.7nF Y2 radial ceramic capacitors, 7.5mm lead pitch **[Kemet C947U472MZVDBA7317]**
12 100nF
1 10nF
1 470pF NP0/C0G
1 100pF NP0/C0G

Resistors *(all SMD M2012/0805 size ⅛W 1% unless noted)*
1 470kΩ
4 100kΩ
2 82kΩ M6332/2512 size 1W **[RC2512FK-0782KL]**
1 18kΩ
1 13kΩ
4 10kΩ
1 5.1kΩ
1 4.7kΩ
3 2.2kΩ
1 2kΩ
10 1kΩ
1 470Ω
3 220Ω
7 12Ω
1 0Ω
1 15mΩ 3W M6432/2512 metal current-sense resistor **[Eaton MSMA2512R0150FGN]**

Optional External Control Box
1 small Jiffy box
1 panel label
3 SPST panel-mount toggle switches
1 1kΩ 16mm potentiometer
1 knob to suit the potentiometer
2 small cable glands
1 1m length of 9-core shielded data cable (or length to suit)

M0+ core running at 64MHz, 32kiB of flash memory and 8kiB of static RAM (SRAM). It includes all the usual peripherals, including a timer designed specifically for motor control applications and comes in a 32-pin 0.8mm-pitch SMD quad package.

CON16 allows IC7 to be reprogrammed in-circuit while CON17 provides a way to power it besides the mains supply.

The motor speed can be set by one of two sources: an external 0-5V control signal or an onboard trimpot. The external speed input enters via pin 2 of terminal block CON8. A 1kΩ series resistor and 100nF capacitor to ground provide noise filtering and protection for the op amp buffer (IC6b). The 470kΩ resistor prevents this input from floating if it is left unconnected.

After buffering, the external speed signal is scaled by the voltage divider formed by the 1kΩ and 2kΩ resistors to suit the 0-3.3V range of the microcontroller's internal analog-to-digital converter (ADC).

The other half of the dual op amp (IC6a) creates a 5V signal to drive one end of the external speed pot. The 5V potential is derived from the 12V rail via the 18kΩ/13kΩ divider and filtered by a 100nF capacitor. It is then applied to op amp IC6a, which is connected as a current-limited unity-gain buffer.

Suppose the current drawn from the 5V terminal is small. In that case, the voltage drop across the 470Ω resistor is low enough that the op amp is not in saturation, and the negative feedback (via the 10kΩ resistor) can maintain the output voltage at 5V.

The op amp output will saturate if the current increases beyond about 14mA with these component values. Voltage regulation will be lost, but the current will be limited to a safe level.

The three digital switch inputs (E-Stop, Run and Reverse) and their respective 12V sources are likewise protected from modest levels of accidental abuse. Taking the E-Stop input at CON9 as an example, the 220Ω series resistor limits the current that can be drawn from the 12V supply.

The signal from pin 2 of CON9 passes through a voltage divider/pulldown/filter formed by 1kΩ and 2.2kΩ resistors plus a 100nF capacitor. Zener diode ZD3 clamps the resulting voltage to a maximum of 5.1V, which is within the safe operating range for the relevant microcontroller I/O pins.

In addition to the external speed control, IC7 has three other analog inputs. The wiper voltages of the internal speed pot VR1 and ramp rate pot VR2 are each fed straight to the micro, with 100nF capacitors providing some noise filtering and buffering for the ADC sample-and-hold capacitor.

The final analog input comes from NTC thermistor NTC2, which monitors heatsink temperature and is connected via CON12. The thermistor forms the upper leg of a voltage divider, with a 4.7kΩ fixed resistor forming the lower leg. The resulting voltage, related to temperature by a non-linear relationship, is fed directly to an ADC channel (PA02 pin 9) on the microcontroller.

The microcontroller drives the two relays and the heatsink fan via more-or-less identical circuits. All three drivers use logic-level Mosfets (Q9, Q10 and Q11) as low-side switches, along with freewheeling diodes (D2, D3 and D4) and 10kΩ gate pulldown resistors. The microcontroller also drives the three LEDs via current-limiting resistors.

The motor control timer inside the MCU uses seven I/O pins – six outputs for the three pairs of PWM signals, plus one input for the fault signal (HOT_FLT). A separate general-purpose I/O pin is used for the enable (PWM_EN) signal. Finally, six digital inputs configured with internal pull-up resistors are used to read the DIP switches (S1).

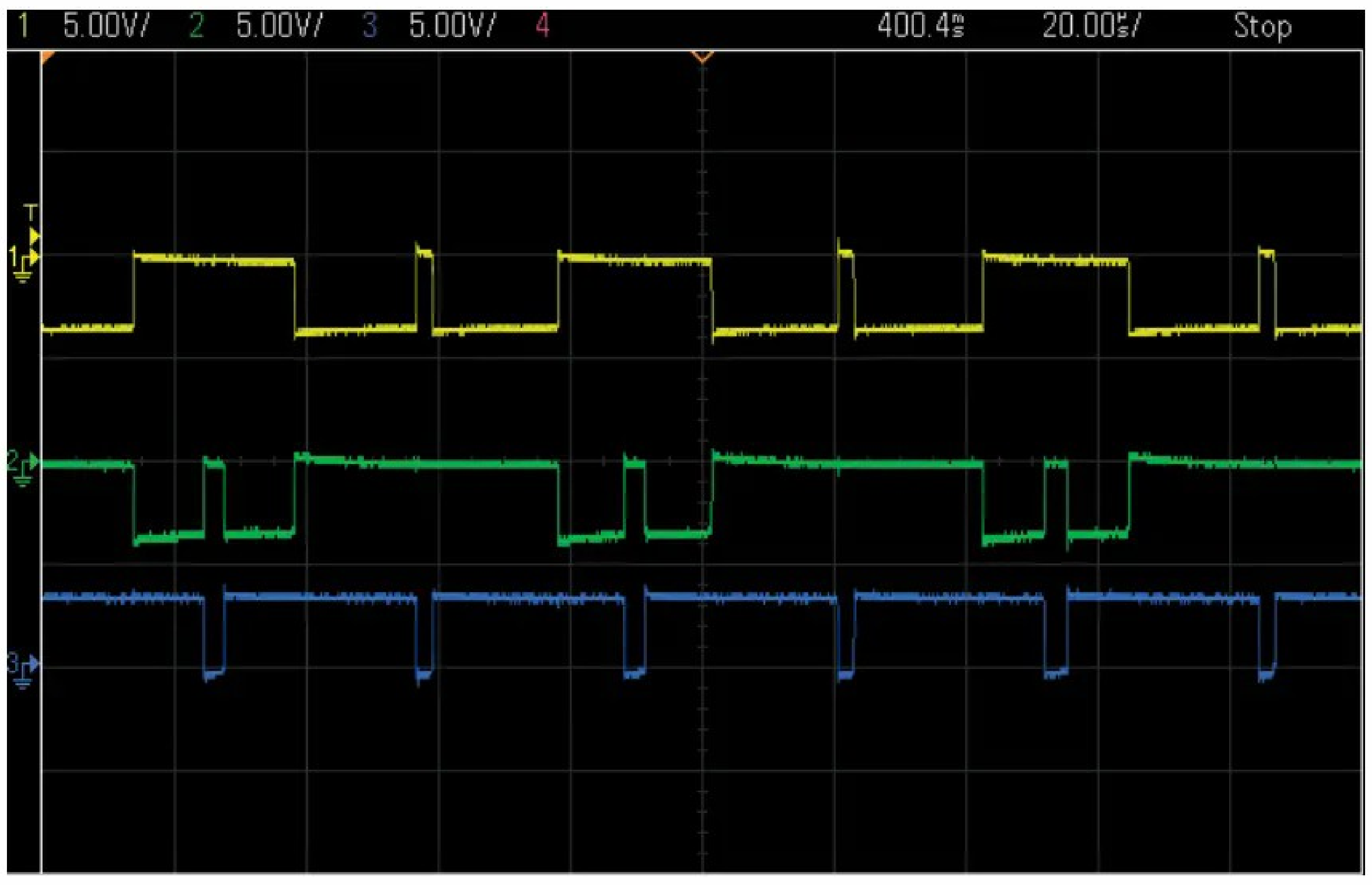
Power for the control circuitry is derived from a second AC-to-DC switch-mode converter module (MOD2), this time one with a 12V output to suit the fan and relay coils. A linear regulator (REG8) derives a 3.3V rail for the microcontroller and associated circuitry from the 12V rail.

Firmware

Of course, a lot of the complexity of a project like this lies in the firmware. Fig.5 shows an overview of the three main blocks of the firmware architecture.

As its name suggests, the I/O driver is responsible for managing all of the I/O functions except those related to the motor-control PWM. On initialisation, this driver reads the mode control DIP switches and stores their values for later use. The driver provides interface functions so the higher-level code can query the state of any switch at any time.

Much of this driver's functionality takes place in a low-priority interrupt service routine (ISR), which is called every 20 milliseconds by a hardware timer. This ISR scans the digital inputs corresponding to the E-Stop, Run and Reverse switches. The inputs are debounced, and the resulting state is stored.



Scope 1: this scope grab shows three traces corresponding to the U, V & W outputs (CON4-CON6). It shows that each is made up of two distinct pulse widths, corresponding to the two phase legs driving it; the use of centre-aligned PWM doubles the effective switching frequency. The vertical scale is 500V/div.

The I/O ISR also starts the sequential analog-to-digital conversion of the four analog inputs (external and internal speed, ramp and heatsink temperature). Direct memory access (DMA) is used to read and store the results when available.

This approach means the reading and processing of the inputs takes place more-or-less automatically. The state machine just has to call an interface function to get the most up-to-date analog or digital input data. In the case of the analog inputs, the reading functions scale the raw ADC values into meaningful units.

The heatsink temperature read function switches the fan on if the heatsink temperature rises above 45°C and off again if it falls below 40°C. If the heatsink temperature exceeds 95C°, an over-temperature error is signalled, and when it drops below 70°C, the over-temperature error is cleared.

Finally, the same ISR manages the flashing of the three LEDs. The state machine code only has to call an interface function once to initiate the flashing of a given LED an arbitrary number of times at a specified rate.

PWM generation

A separate module looks after the generation of the motor PWM signals. The timer used to generate the PWM includes (among many other things) a 16-bit counter and three comparison

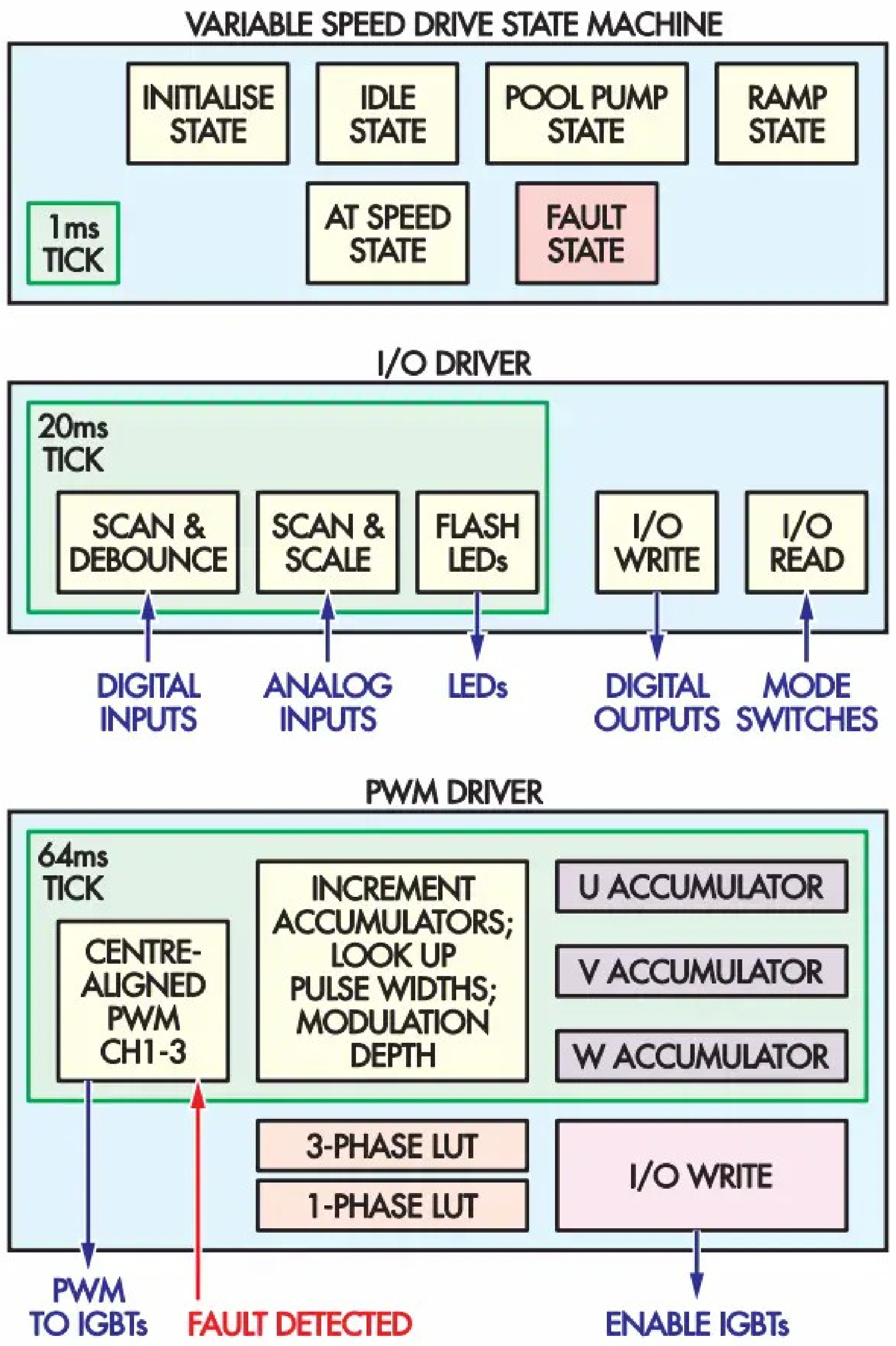

VARIABLE SPEED DRIVE STATE MACHINE
INITIALISE STATE
IDLE STATE
POOL PUMP STATE
RAMP STATE
1ms TICK
AT SPEED STATE
FAULT STATE
I/O DRIVER
20ms TICK
SCAN & DEBOUNCE
SCAN & SCALE
FLASH LEDs
I/O WRITE
I/O READ
DIGITAL INPUTS
ANALOG INPUTS
LEDs
DIGITAL OUTPUTS
MODE SWITCHES
PWM DRIVER
64ms TICK
CENTRE-ALIGNED PWM CH1-3
INCREMENT ACCUMULATORS; LOOK UP PULSE WIDTHS; MODULATION DEPTH
U ACCUMULATOR
V ACCUMULATOR
W ACCUMULATOR
3-PHASE LUT
1-PHASE LUT
I/O WRITE
PWM TO IGBTs
FAULT DETECTED
ENABLE IGBTs

Fig.5: the firmware's three principal blocks. An I/O driver manages the digital and analog interfaces, a PWM driver generates the motor control signals, while a state machine controls the overall system logic.

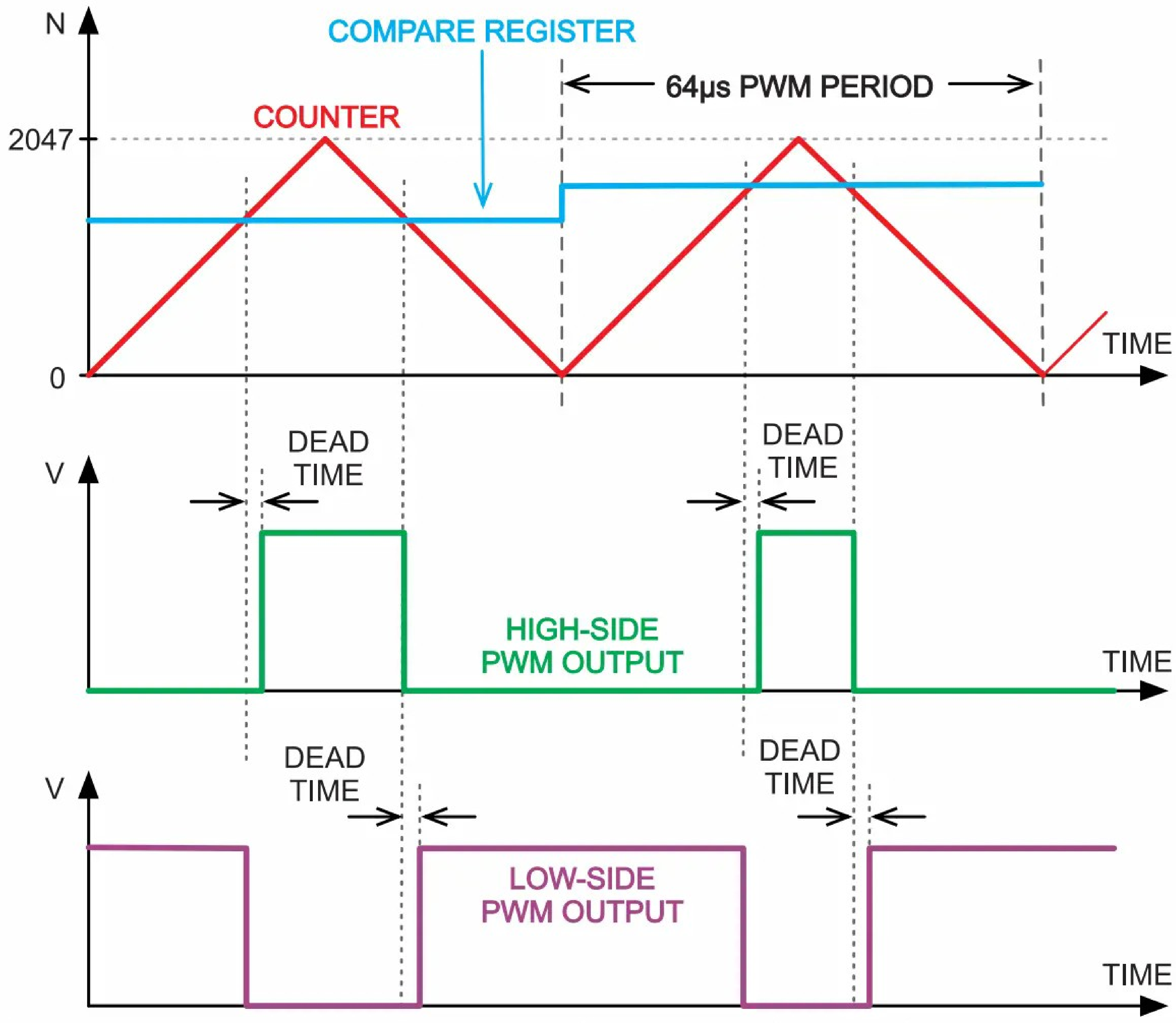

N
COMPARE REGISTER
64µs PWM PERIOD
COUNTER
2047
0
TIME
V
DEAD TIME
DEAD TIME
HIGH-SIDE PWM OUTPUT
TIME
V
DEAD TIME
DEAD TIME
LOW-SIDE PWM OUTPUT
TIME

Fig.6: centre-aligned PWM is preferred for motor drive applications since the switching edges of each phase are not aligned, doubling the effective switching frequency seen by the motor windings and reducing EMI/RFI.

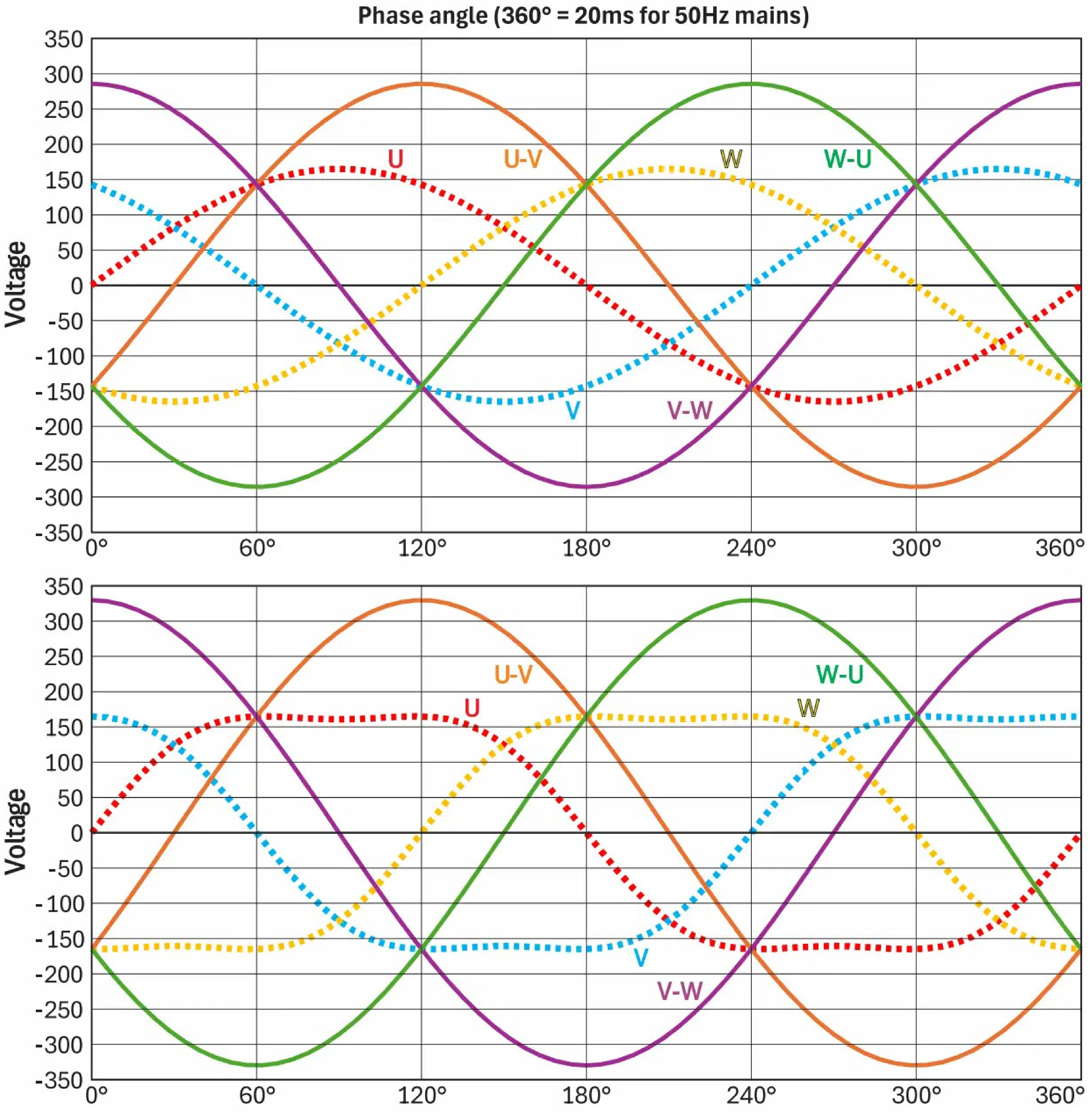


Fig.7: if we modulated each phase with a pure sinewave, the phase-to-phase output voltage would only be about 87% of the maximum (at top). Adding third harmonic content to the modulation allows us to achieve the maximum phase-to-phase voltage (around 230V RMS), demonstrated in the lower plot.

registers. The counter is clocked at 64MHz and is programmed to count from zero up to 2047, then down again to zero, as shown diagrammatically in Fig.6.

On every clock cycle, the counter value is compared to the value in the compare registers to generate a centre-aligned PWM signal, as shown in that figure. Centre-aligned PWM is preferred for motor control since the switching edges on each phase are not aligned with each other, as would be the case if edge-aligned PWM was used.

This means the phase-to-phase voltage across the motor windings switches twice as often, doubling the effective switching frequency and PWM resolution.

The phase (IGBT) switching frequency is 15.625kHz, but the motor phase-to-phase windings see switching at twice this rate, or 31.25kHz, as you can see in Scope 1. This shows the three phase-to-phase voltages at a scale of 500V per division.

You can see that each waveform has two different pulse widths, corresponding to the phase legs driving each end of the winding. The result is two transitions each 64μs period.

The motor control timer also takes care of generating the complementary output signals to drive the high-side and low-side switches and inserting a dead-time between them, as shown in Fig.6.

The timer's final job is to ensure the outputs are placed in a known state if there is a fault. In our case, the timer is configured to switch them all low, turning off all the IGBTs, although this is fully configurable.

This leaves our PWM code with the task of loading an updated pulse width value into each compare register every 64μs PWM cycle. To do this, a 32-bit 'accumulator' for each phase is incremented each time by an amount proportional to the desired output frequency.

The upper eight bits of the accumulator are used as an index into a look-up table containing 256 samples of one cycle of the output waveform we want to produce. The appropriate sample is extracted, scaled according to the required output voltage, and loaded into the relevant compare register.

Two accumulators and two PWM channels are used for a single-phase motor. The accumulators are initialised to values representing 0° and 180° in the table. The table contains values representing a sinusoid.

For three-phase operation, three accumulators and three PWM channels are used, with the U, V and W accumulators initialised to positions 0°, 120° and 240° into the table for forward rotation or 0°, 240° and 120° for reverse rotation.

Unlike the single-phase look-up table, the three-phase table does not contain samples of a pure sinewave. Instead, it contains values representing a sinusoid with about 16% of added third harmonic. Fig.7 shows why this is necessary.

Starting at the top, sinusoidal phase voltages with a peak-to-peak value of 330V (shown dotted) produce phase-to-phase voltages (solid lines) with a peak-to-peak value of 570V. This corresponds to an RMS voltage of just 200V RMS, not the 230V we desire.

If we modulate the phase voltages with a sinewave with an added third harmonic, as shown below, the peak-to-peak phase voltages are the same as before, but the wave shape is very different. The resulting phase-to-phase voltages are nonetheless sinusoidal, but their peak-to-peak value is now 660V, giving an RMS voltage of 230V.

State machine

With the I/O and PWM taken care of, all that remains is to implement the motor controller's application logic. This is done using a simple state machine. A state machine (properly a finite state machine) is a computational model that can be used to implement complex behaviour in a structured manner.

The behaviour is modelled by several states, only one of which can be active at any given time; a set of transition rules determines how and when the machine can transition from one state according to external trigger events. Each state can have actions that are executed when it is entered, exited, or when a trigger event occurs.

The simple version used here is always triggered by a regular timer 'tick', prompting the state machine to

Table 1: Software States

	Entry Action	Trigger Action (Transition Rules)	Exit Action
Initalise state	- Initialise internal variables - Start IO Driver (reads mode switches) - Start PWM Driver (specify 1-phase or 3-phase) - Flash red, yellow & green LEDs twice, fast - Start soft start bypass timer (3 seconds)	**- if soft start bypass timer expired:** - if 1-phase & Pool-mode transition to Pool-Pump state - else transition to Idle state - else no transtion	- Close soft start bypass relay
Pool Pump state	- Flash green LED slowly - Start pool pump timer (30 or 300 seconds) - Set speed_now to zero - Enable PWM	- if fault transition to Fault state - if E-Stop open transition to Idle state **- if Run open:** - if speed_now > min_speed transition to Ramp state - else transition to Idle state - if pool pump timer expired transtion to Ramp state - if speed_now < pool_pump_speed increment speed_now - else no transition	- Turn green LED off
Idle state	- Disable PWM - Set speed_now to zero - Turn yellow LED on - Start idle dwell timer	- if fault transition to Fault state - if idle dwell timer running no transition - if E-Stop open no transition - if speed_req > min_speed transition to ramp state	- Turn yellow LED off - Read Reverse pin state
Ramp state	- Flash green LED indefinitely fast - Set PWM direction (ignored if 1-Phase) - Set PWM speed to speed_now - Enable PWM (ignored if already enabled)	- if fault transition to Fault state - if E-Stop open transition to Idle state - Get speed_req (speed demand, Run & Reverse states) **- if speed_now ≤ speed_req – margin:** - Increment speed_now (based on ramp, limit to speed_req) - Set PWM speed to speed_now, no transtion **- else if speed_now ≥ speed_req + margin:** - Decrement speed_now (based on ramp, limit to speed_req) - if speed_now < min_speed transition to Idle_state - Set PWM speed to speed_now, no transition - else transition to At-Speed state	- Turn green LED off
At-Speed state	- Assert At_Speed output (ignored if not enabled) - Turn green LED on	- if fault transition to Fault state - if E-Stop open transition to Idle state - if speed_now ≤speed_req – margin transition to Ramp state - if speed_now ≥ speed_req + margin transition to Ramp state - else no transition	- Deassert At_Speed output - Turn off green LED
Fault state	- Disable PWM - Set PWM speed to zero - Clear E-stop cycle flag - Assert Fault output (ignored if not enabled) - Set red LED - Set yellow LED if overtemp fault	**- if faults cleared:** **- if E-Stop cycle flag clear:** - if E-Stop open set E-stop cycle flag - no transition - else if E-Stop closed transition to Idle State - else no transition - else no transition	- Turn off red & yellow LEDs - Deassert Fault output

assess the transition rules associated with the current state and initiate a transition if required.

If a state change is required, the state machine executes the current state's exit actions, switches to the new state and executes its entry actions. The following trigger causes the new state's transition rules to be evaluated.

States are defined by three functions: an entry function containing the entry actions; a tick function containing the state change rules and tick actions; and an exit function containing the exit actions. By partitioning the VSD's operation in this way, the controller's logic becomes easier to understand and therefore implement and maintain.

You can see this in Table 1, which describes the VSD's operation in one neat summary. A total of six states are required, including an initialisation state where execution starts. The timers described in the table are software timers driven by a 1ms interrupt provided to the state machine. The 'tick' time in the VSD is set to 100ms, meaning the state transition rules are evaluated 10 times per second.

Conclusion

That's all we have room for this month. Next month, we will cover the construction, testing and use of the VSD.

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Note '10/63' denotes '10µF 63V'.

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Contact

Jake Rothman
Manchester House
Market Street, Craven Arms
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AUDIO OUT

By Jake Rothman

The Transmanium fuzz box, part 2 – assembly

Last month, I described the circuit design and PCB for this new fuzz box, based on germanium transistors. Now it's time to put it all together, giving the result shown in Photo 1.

Being a foot pedal for musicians, the criteria are different from a normal desktop electronic enclosure. It has to be tough, because it's on the floor and the effect is switched in/out by a brutal boot on what's literally called a stomp switch. A metal enclosure is needed; plastic will only last a few weeks in use.

It is also necessary for screening, with it being a high-gain, high-impedance unit. The case is Earthed through the pot bushes via the PCB. If the case is painted, it will be necessary to de-burr the hole edges to bare metal so that the pot bodies make electric contact. If even the switch body isn't Earthed, clicks will result.

The knobs have to be top-mounted so that you can see the settings, and easily make adjustments which makes the pot shafts vulnerable. As a result, Solid metal shafts are preferred.

If plastic shafts are unavoidable, a 6mm diameter is the minimum. The knobs have to be spaced away from the switch area by a least 40mm to avoid damage. This necessitates a relatively long, large box for the electronics contained within.

It is tradition and good practice to have the sockets on the back to keep the leads out of the way of the performer. Photo 3 shows the layout of the mounted components.

Rubber hole plugs have been fitted as 'feet' to stop the unit from moving about on the floor, visible in Photo 2.

Most guitarists stick their pedals to a pedalboard with Velcro, so these can be left off if desired. It saves a lot of drilling.

Boxing clever

Most pedal manufacturers have to invest a considerable amount into enclosure tooling. The Colorsound presses cost well over £5k. Luckily, they survived the Hemel Hempstead Buncefield oil explosion in 2005, which destroyed the factory. For small-scale production, ready-made die-cast enclosures are the best because they are easy to drill.

Companies like Hammond Manufacturing and Tayda make excellent cases in many colours and offer a small-quantity drilling service.

The drilling diagram can be downloaded from https://pemag.au/link/ac8p. Ready-drilled boxes will be made available, which will be used as a format for further designs as well.

Readers can have the Tayda drilling file on request. Because the sides of diecast boxes are sloped with curved edges (to enable release from the mould), it is difficult to give accurate measurements from the edge. The way around this is to measure from the

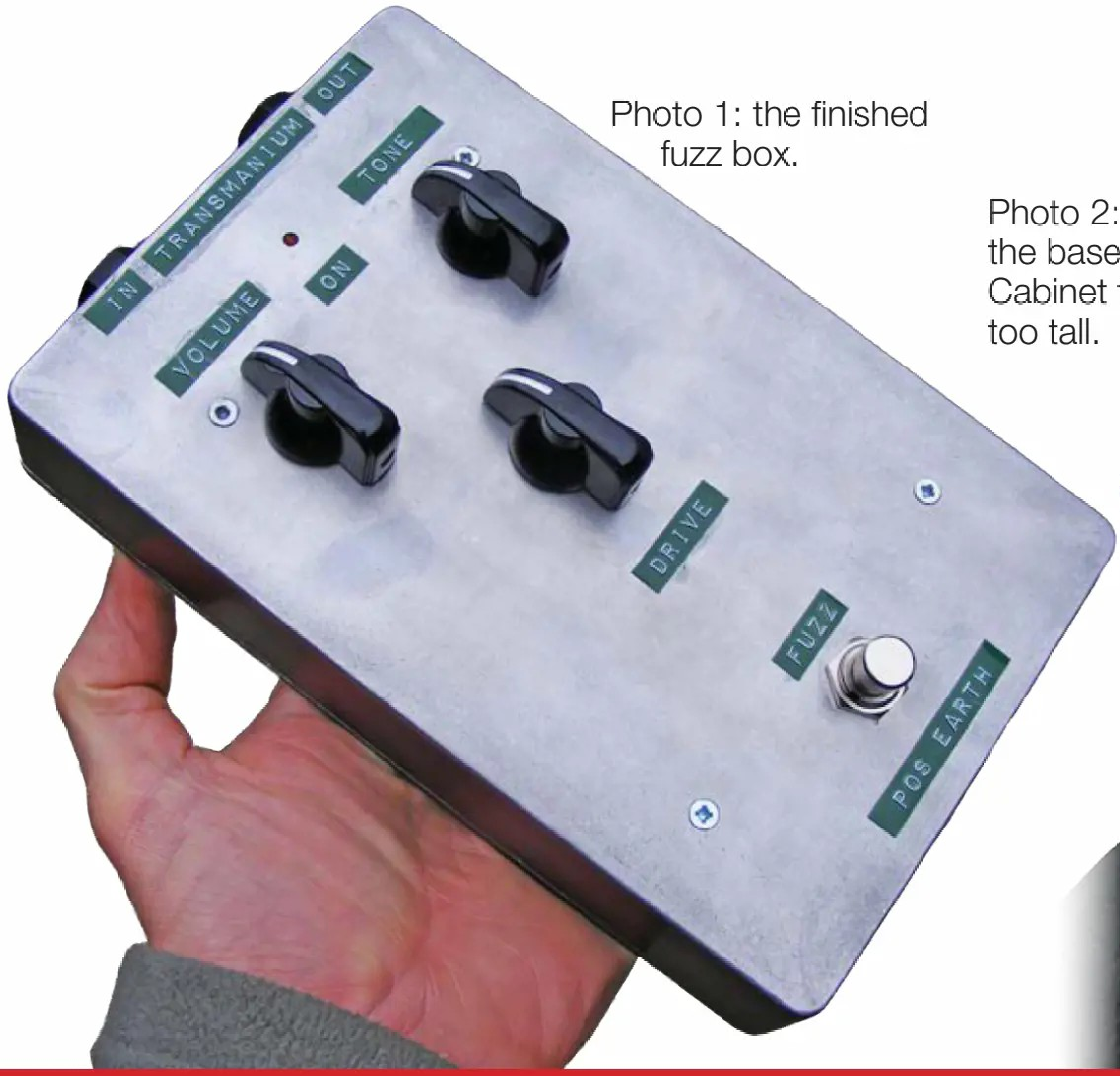

Photo 1: the finished fuzz box.

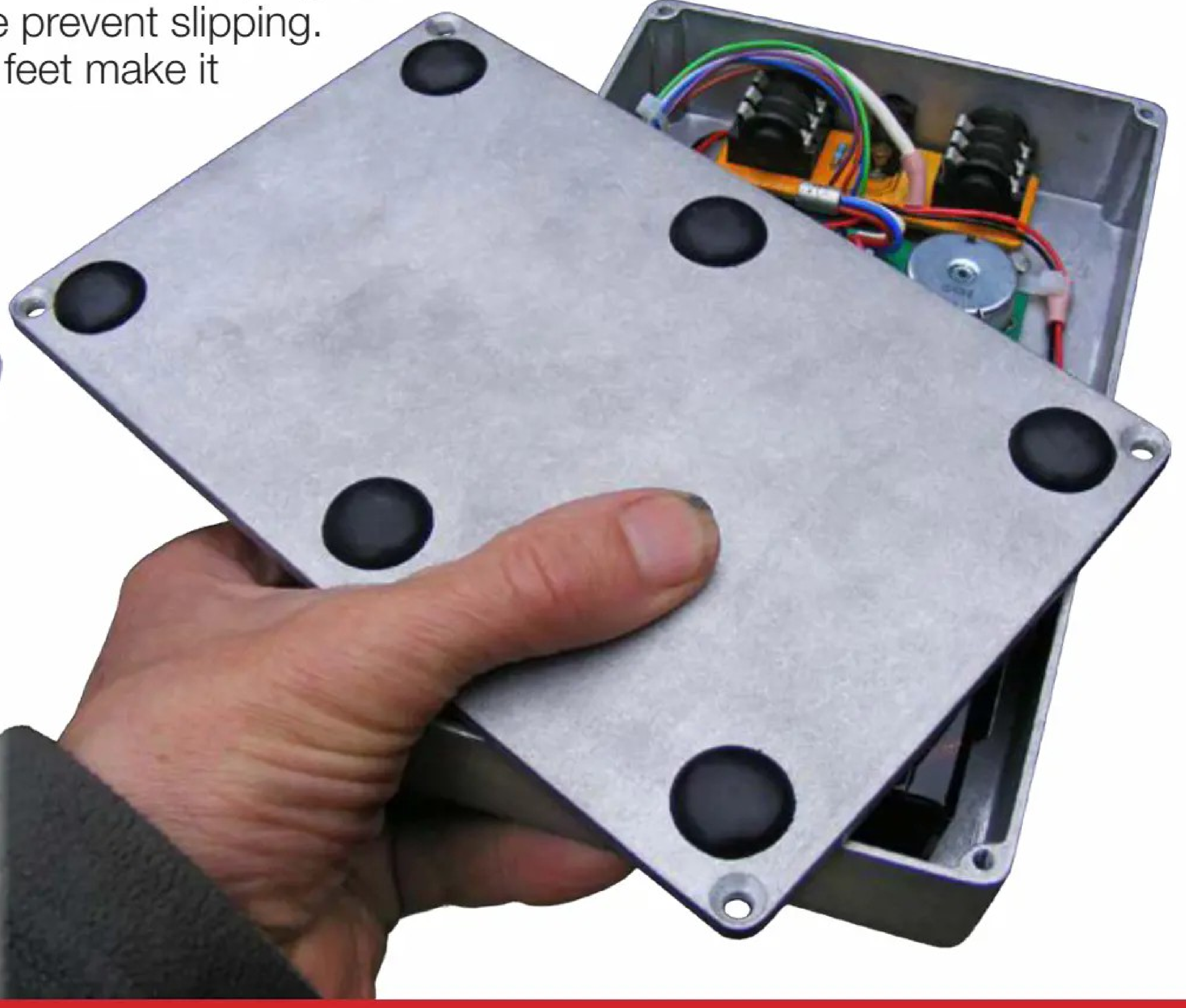

Photo 2: rubber hole plugs on the base prevent slipping. Cabinet feet make it too tall.

Photo 5: The back of the box with the jacks. The plastic plate hides the wonky hole for the barrel socket.

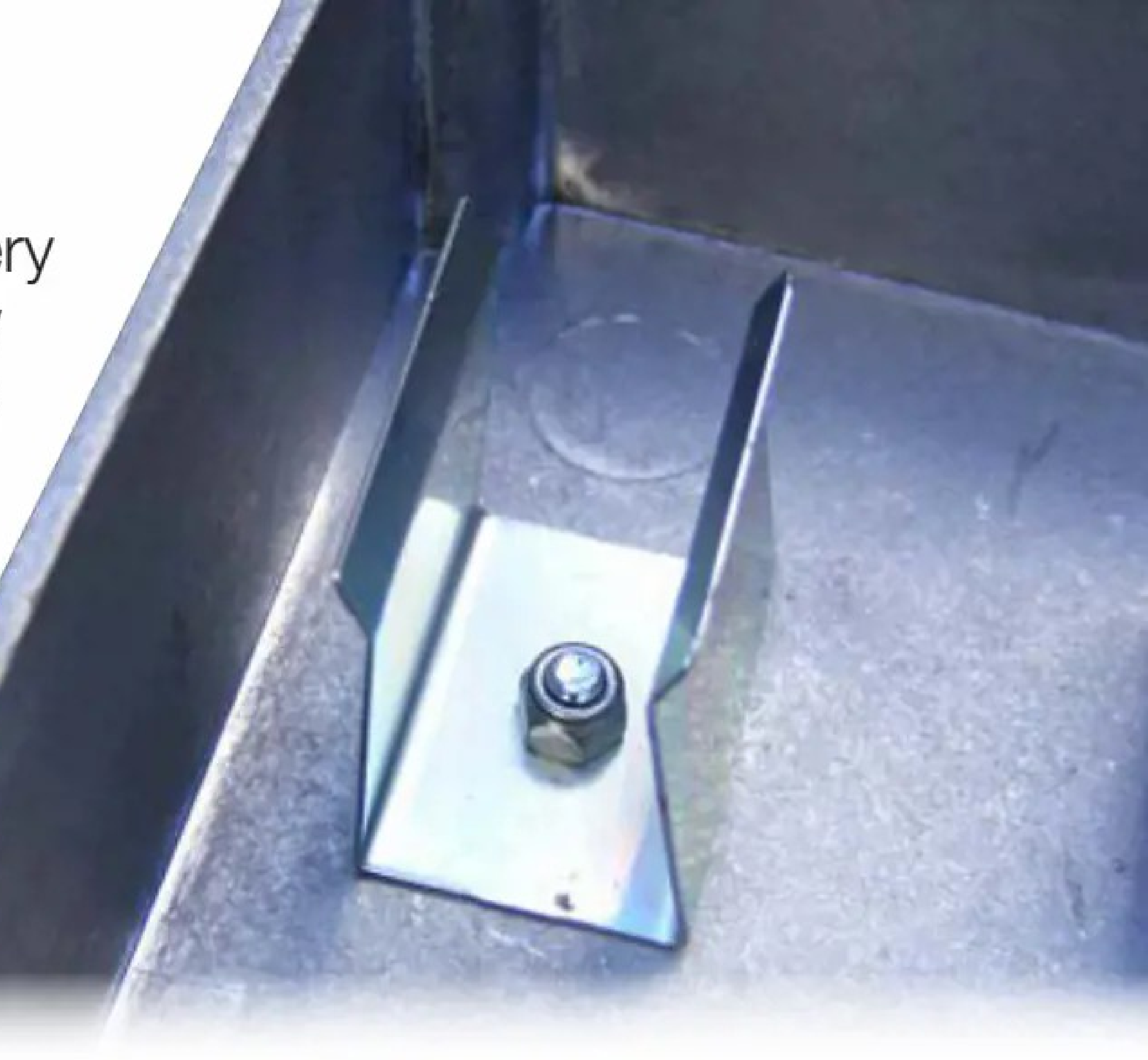

Photo 6: The battery should be securely fitted in a clip. Use a Nylock nut so it doesn't back off. It's not good enough to just use a lump of foam like some pedals.

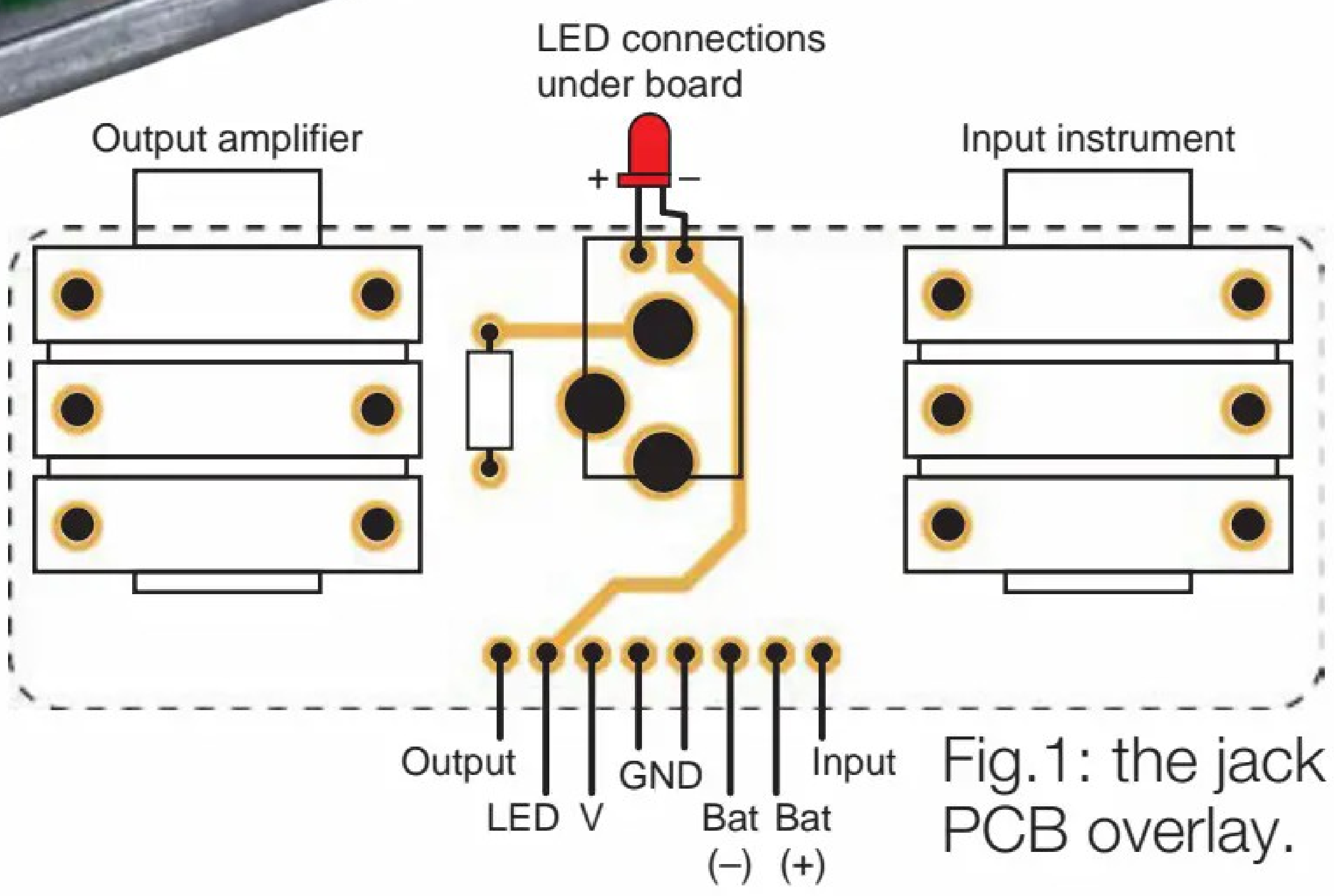

Fig.1: the jack PCB overlay.

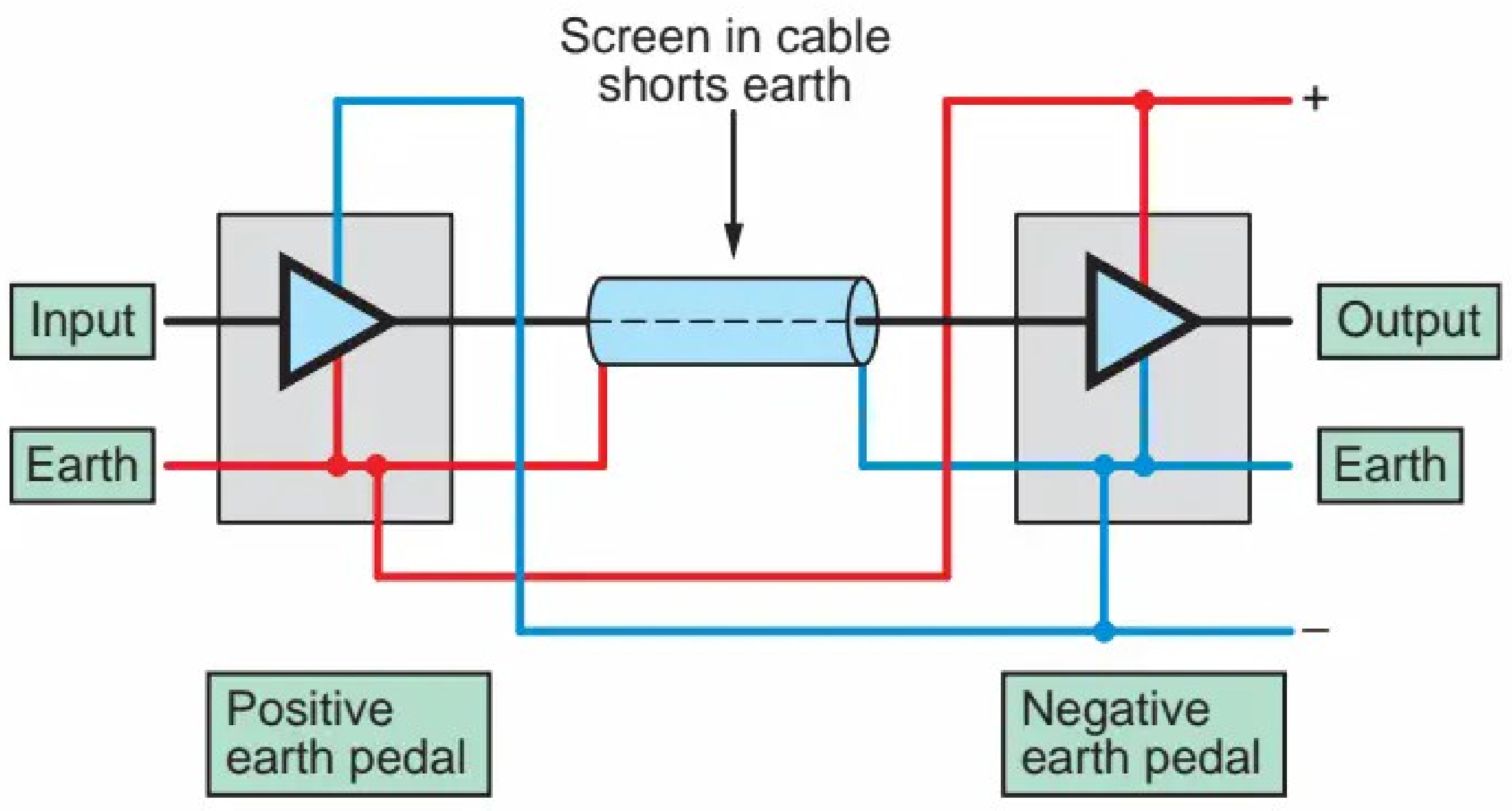

Fig.2: when a positive Earth pedal is powered by the same supply as a negative Earth unit, it is shorted out.

centre lines of the box, as Tayda specifies for their drilling.

Jack board

The 2° ±1° sloped sides also mean it is difficult to mount the jack sockets on the main board without stressing the pins. This problem is avoided by using a separate jack board, shown in Photo 4. Fig.1 is the overlay diagram for this board.

Of course, the sockets can be hard-wired without this board if that's how you want to roll. This makes assembly more difficult, but future parts replacement easier. I have found PCB-mounted 6.3mm jacks to be very reliable. The 5.5mm outer diameter, 2.1mm inner diameter power barrel jacks are not so good.

On this unit, the sockets follow normal circuit signal flow convention with the input on the left and the output on the right when looking at the front panel. This is how Colorsound do it, but I've since found most guitar pedals do it the other way around, because the guitarist's lead comes out on the right (at least, if they're a right-hander playing a right-handed guitar).

Having the pedal's input on the right avoids crossing the leads over. I will make a 'normal' board option later; it could be done simply by turning the board over, but this means the components can't be seen, an aesthetic no-no in guitar pedal circles. Also, the power socket hole would have to be repositioned after flipping it over.

The socket end is shown in Photo 5. Note how the power socket sticks out from the board slightly (Photo 4). This necessitates the use of washers on the inside of the audio jacks.

Positive Earth

Because the circuit uses a positive Earth, the power socket is centre-pin positive, which goes against the guitar pedal convention. It is necessary for this pedal to have its own power supply to prevent shorting with negative-Earth pedals anyway, so I don't think that is a problem. This shorting is illustrated in Fig.2.

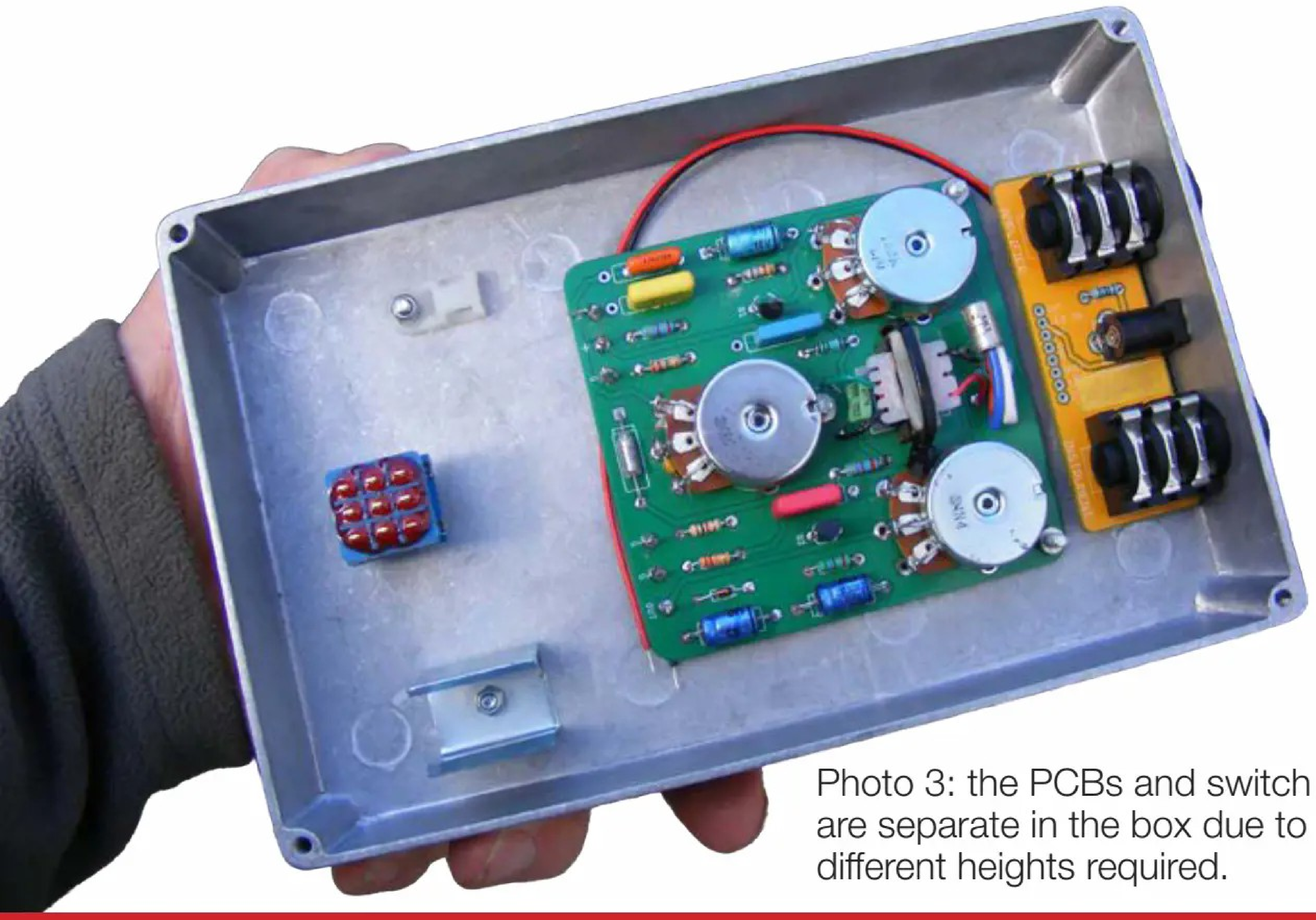

Photo 3: the PCBs and switch are separate in the box due to the different heights required.

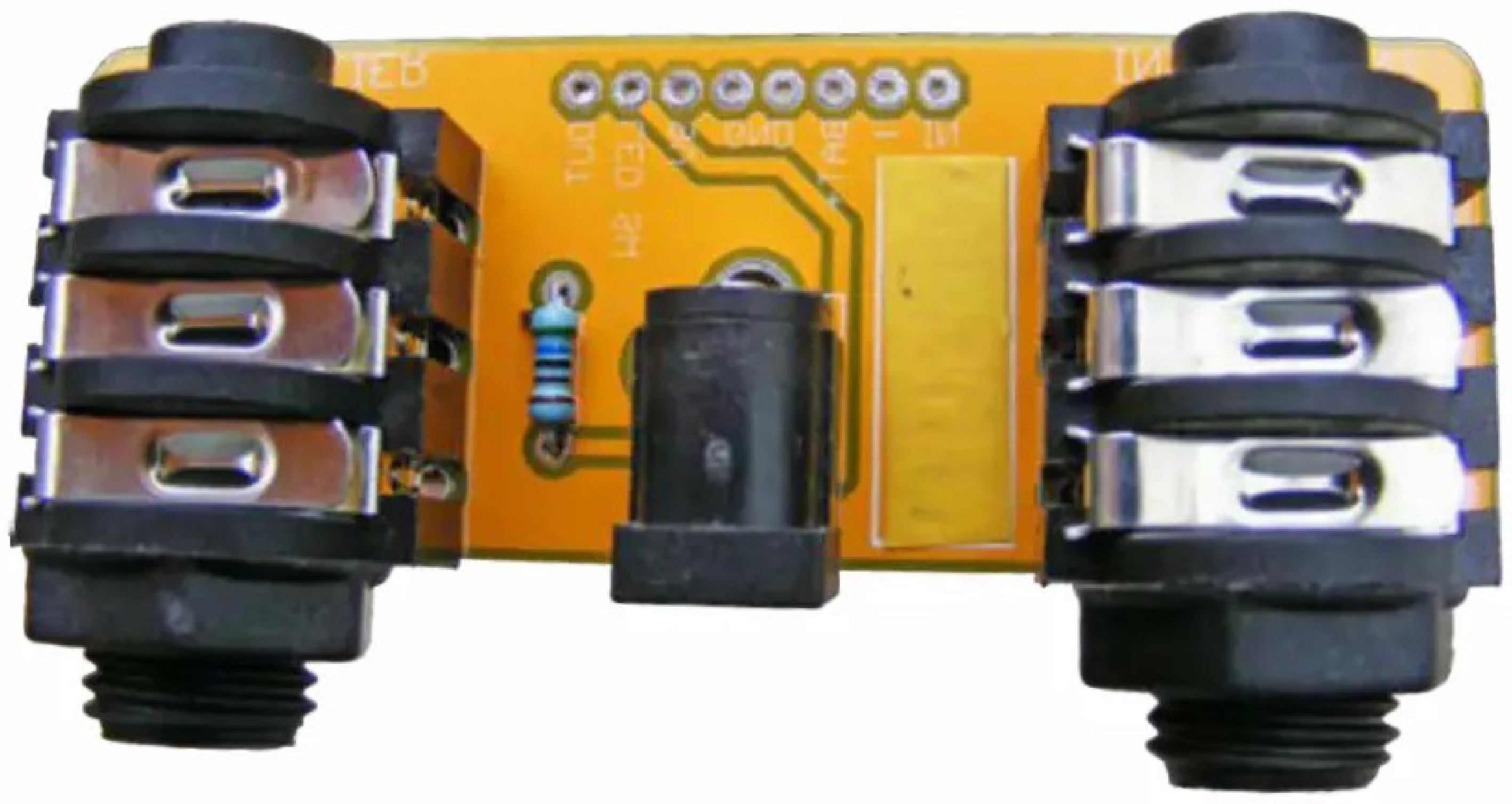

Photo 4: the jack PCB. Remember to put the jack washers on before you mount it in the box.

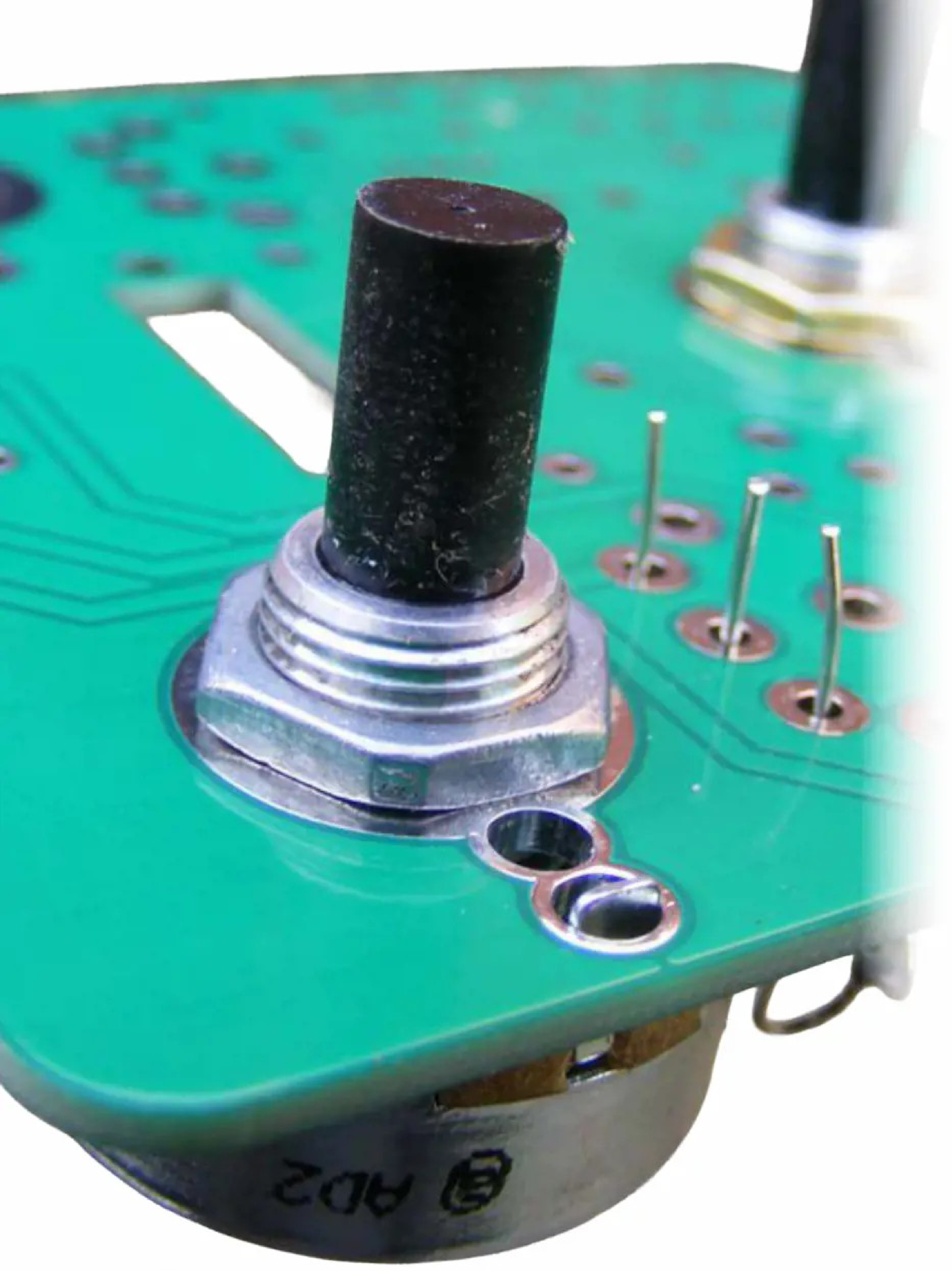

Photo 7: these Alpha 24mm pots from Rapid had insufficient bushing length to reach the case so they are secure to the board. Always solder after tightening to prevent stress on the joints.

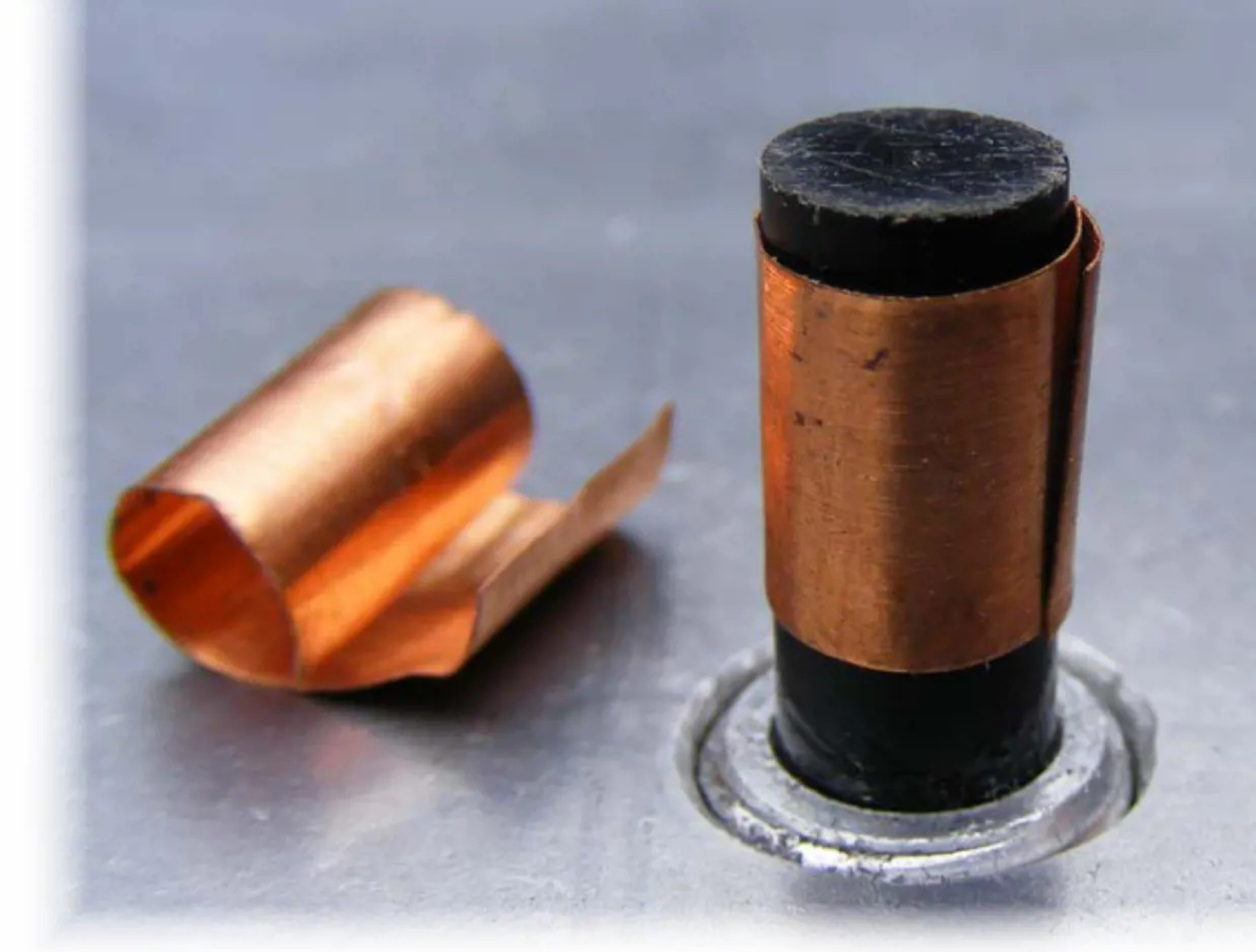

Photo 8: a thin copper foil shim can be wrapped around the 6mm shaft to fit 6.35mm (¼-inch) knobs.

The current consumption is 3.7mA, so an ordinary zinc-carbon PP3-type battery lasts a long time. The battery needs to be held securely by a clip (Rapid) as shown in Photo 6. Use the best battery connector ('snap' to guitarists) you can find. The ones that are moulded are less likely to fail than the crimped plastic-film ones.

Power switch?

Another strange convention with guitar pedals is that no power switch is provided. Power switching is achieved by using a jack socket as a switch, switching on the unit when a jack plug is inserted.

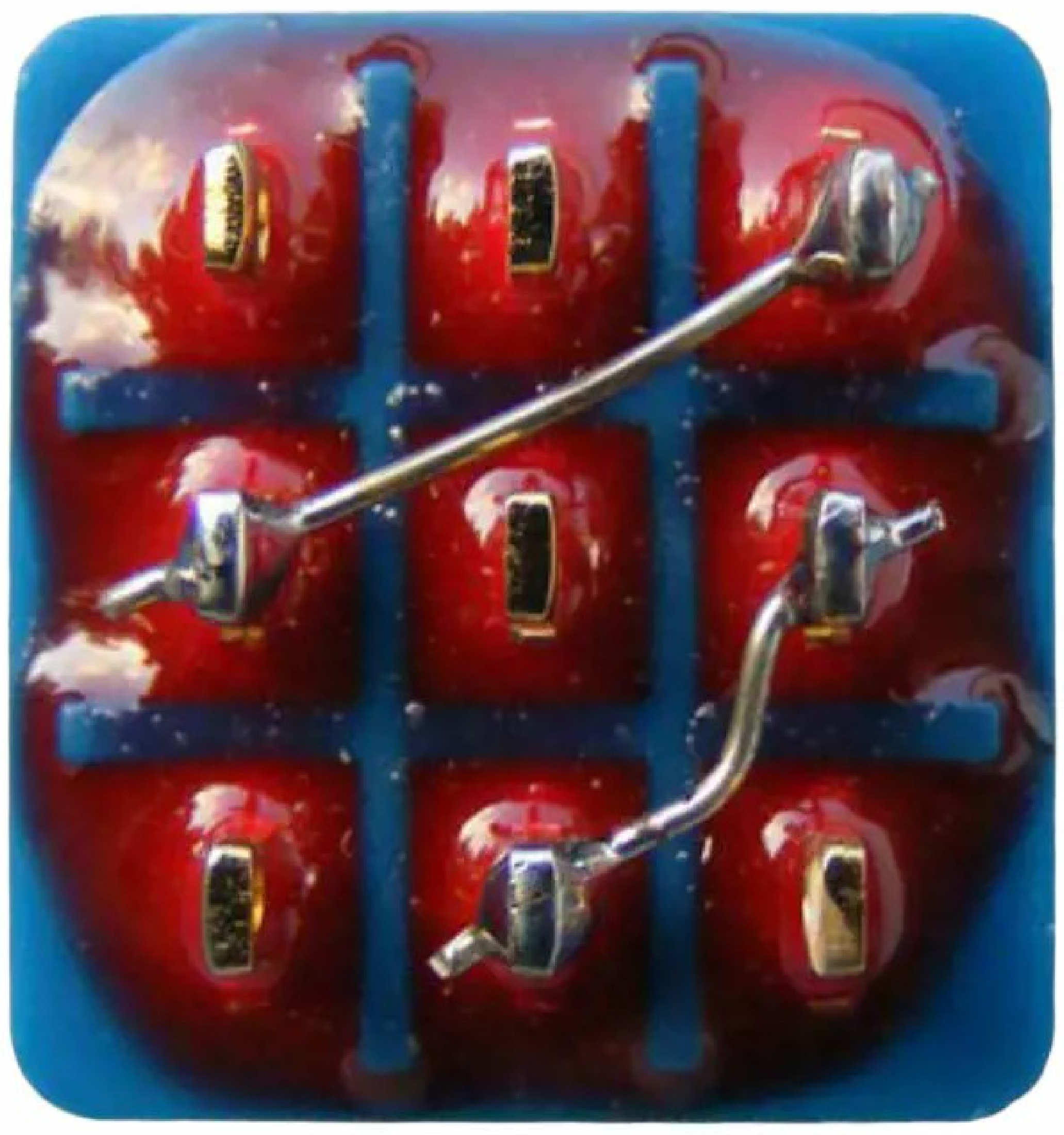

Photo 11: wiring the switch links.

This is achieved by the long Earth sleeve of a mono jack plug shorting together the sleeve and ring contacts of a stereo jack socket (or using a socket with a built-in switch actuated by the plug shaft).

The Earth connection of the battery is connected to the centre terminal/ring of the jack. Thus, it is the Earth connection that is switched, not V_{CC}.

It is best to perform the switching on the input jack to minimise the loud crack that occurs when the lead is inserted.

Although guitarists don't seem to care about loud bangs, I think the amplifier should be muted when plugging in to protect one's hearing. Lastly, don't forget to pull the lead out when you are finished, or you may end up with a flat battery.

Pot problems

An unforeseen problem with the diecast case was its thickness compared to the pressed steel Colorsound cases, which meant the bushings on most pots were not long enough to mount to the PCB (see Photo 7).

The solution, for now, is to fix the board with mounting holes using countersunk head M3 screws. An expensive solution is to order Omeg or Bourns pots with the longer bushings.

The final solution will be milled holes in the PCB to accommodate normal cheap 16mm pots with standard pins. My PCB designer, Mike Grindle, is working on it now.

Shafted

Another pot irritant is shaft size. Most low-cost pots with round shafts have a 6mm diameter, while most knobs are for ¼-inch or 6.35mm diameter: a typical engineering imperial/metric mix-up. This difference causes an ugly eccentricity as the knob is rotated.

One way around this is to add a thin copper strip shim around the shaft (as in Photo 8), or better still, buy some brass adaptor tubes from the synthesiser shop Thonk or Tayda, as shown in Photo 9.

Knobs

I don't like push-on knobs because they often fall off. They are also generally designed to fit splined shafts that are prone to snapping because the split weakens the structure. They are great for manufacturing though, where time is money.

The home constructor can use quality grub-screw fixing types, although there is a subtle art to mounting them. Having fitted thousands in my time, I make it look easy, but it is not.

The key is to turn all the shafts hard one way and then put the knobs on. Lift the knobs by 0.5mm and line up the marker line/pointer and tighten the screw a little bit, almost at the same time.

Turn the knob fully the other way to make sure it's symmetrical and not scraping. If it's all okay, fully tighten the screw. Occasionally, grub screws are lost, rendering the knob useless.

No need to worry, they are now

Photo 9: ready-made pot shaft adaptors are an even better solution.

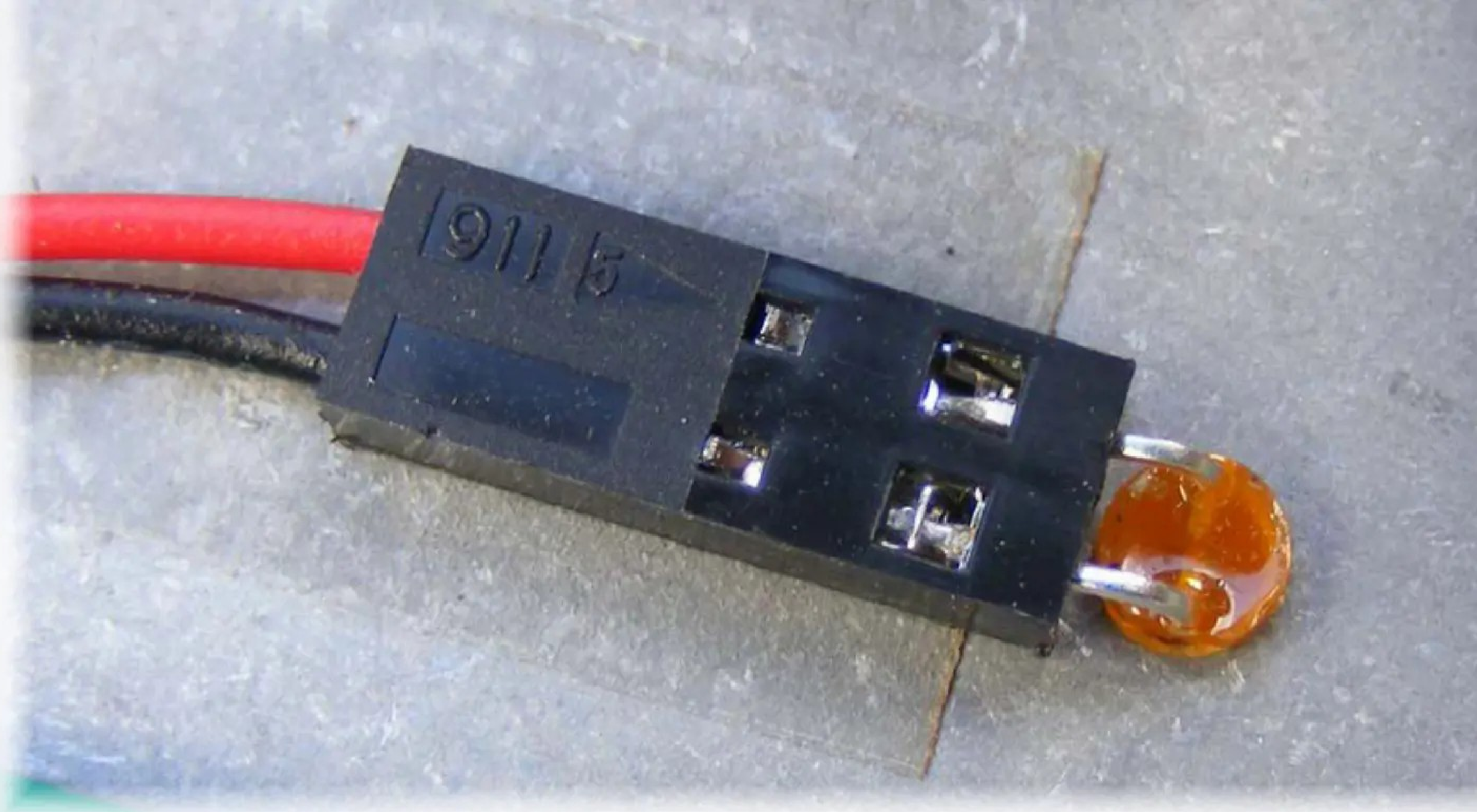

Photo 10: the LED is mounted using a solderless connector under the jack board. A bit of Evostik (rubber-based glue) will help keep it in place.

available from Tayda for 3p each following a lot of pestering. If the screw is obscure, such as a 4BA type, I've found model suppliers have them, but they are expensive.

For pedals, I like to use traditional pointer (aka 'chicken head') knobs as shown or 'Marshall-style' silver-cap types with marker lines. These are available from Tayda and Action Hardware.

Bypass switching

The fuzz box uses a 'hard' bypass that enables the guitarist to carry on playing if there is a failure, such as a flat battery. The circuit is shown in Fig.3 overleaf. The input of the fuzz box is grounded to prevent hum pickup when bypassed. This helps prevent a click when the effect is switched back in.

LED wiring

A switch section is also used for the LED. A 5.6kΩ current-limiting resistor (marked CLR) gives sufficient brightness with high-efficiency LEDs, such as the Tayda or Kingbright types shown. I always prefer orange LEDs myself; blue ones give me 'mental health issues'!

I tried to mount the LED directly underneath the PCB, but the convoluted bending caused too much stress and cracked the LED's case. It's a tricky PCB design problem because the LED needs to go where the power socket pads are located.

For now, until Grindle's spatial brain finds a solution, a solderless LED holder (Rapid or Tayda) is used, stuck down with double-sided tape, as shown in Photo 10.

Remaining wiring

The next job is to put the links on the switch (see Photo 11) using 24 SWG (0.56mm diameter) tinned copper wire or resistor lead off-cuts.

The interconnecting wires between the switch and the two boards are shown in Photos 12 &

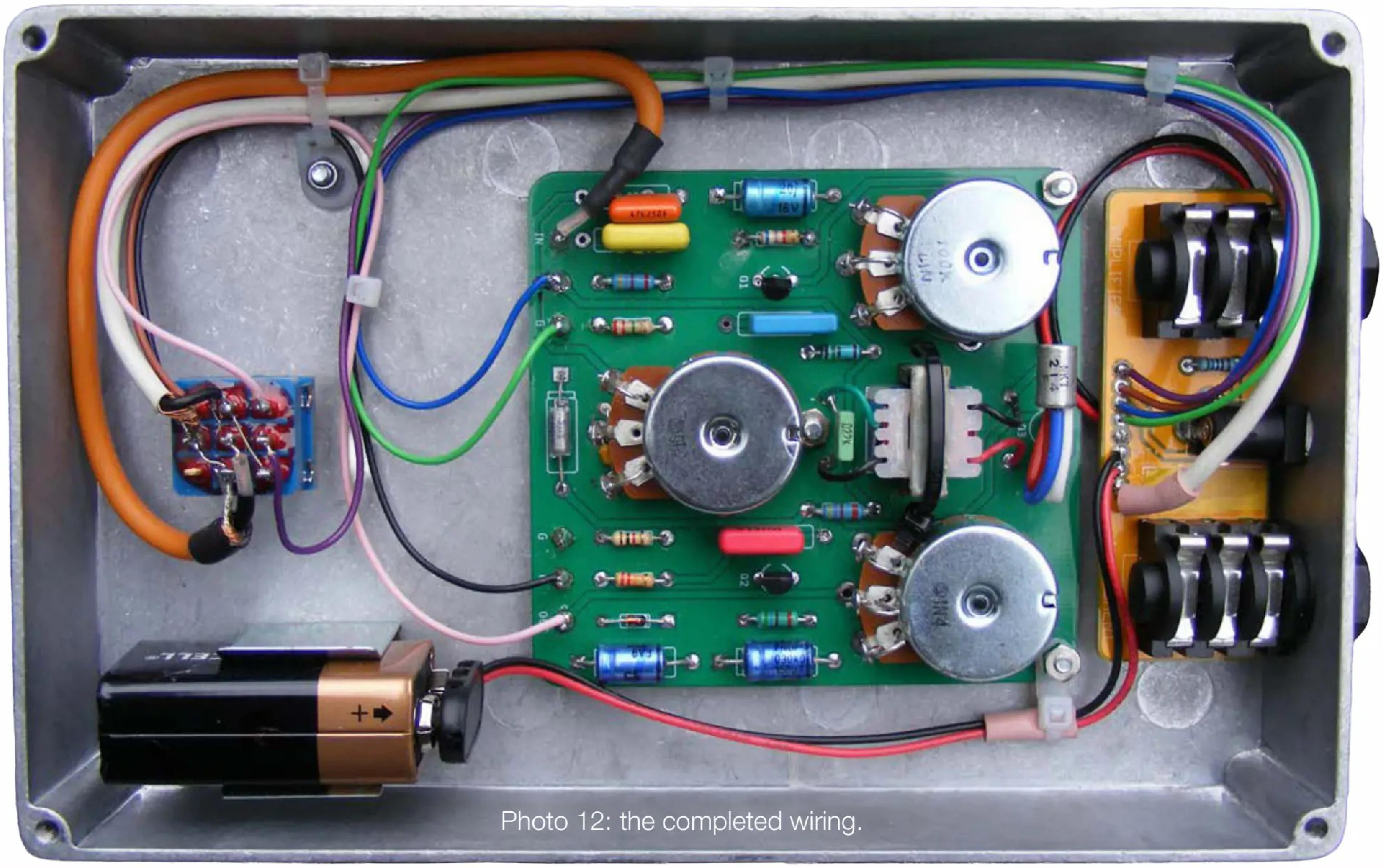

Photo 12: the completed wiring.

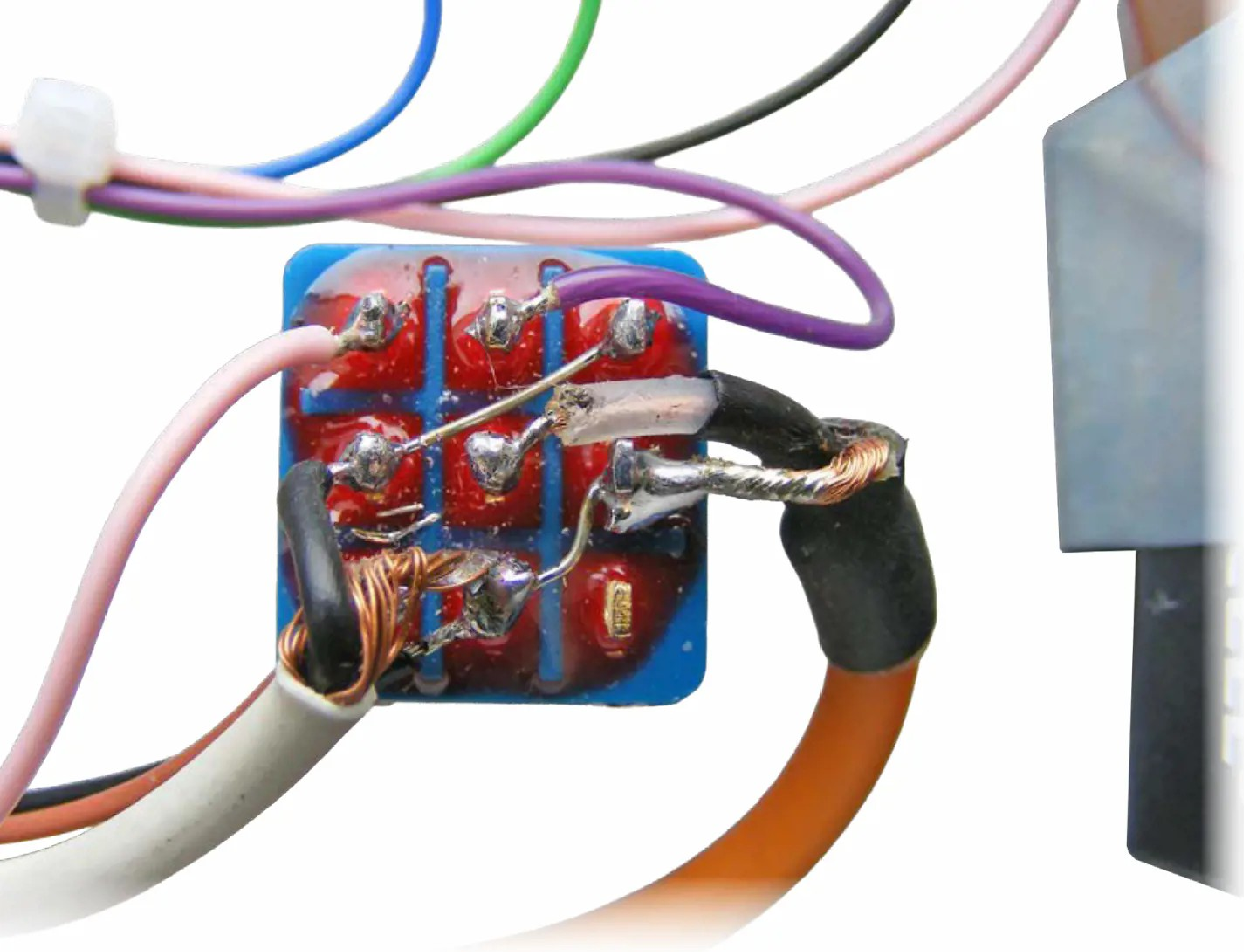

Photo 13: the switch wiring is a difficult job. The cable screens go to the earth link on the lower right hand corner and are not connected at the other end.

13. I always use 7/0.2 stranded wire for this sort of work. Solid-core wire snaps off with vibration unless very well clamped.

The completed wiring loom is shown in Photo 12. There are two screw-mounted cable clamps. One is for the battery clip wires, with a bit of silicone rubber sleeve for grip, with the other for the main loom.

Clicks can be reduced by using a screened wire for the input cables, since ordinary wire becomes microphonic in high impedance situations, generating a voltage when it moves, such as when the switch is pressed.

Because the sensitive input cables are screened, all the wires can be cable-tied together without feedback occurring. The switch wiring is the most difficult part of the construction, so great care is needed (see Photo 13).

Don't overheat the tags or move the wire while soldering. The plastic can melt, losing the internal connection, rendering the second most expensive part useless.

Testing

At least a simple thing like this can be fully tested, unlike modern software-based equipment. After all, there are only three knobs!

The most common problem is hum and instability from poor Earthing. There is a provision for the stabilising capacitor (C9) and tuning capacitor (C10) on the new boards being supplied, if required.

Be aware that with this style of construction, it is possible for shorts to occur from the underside of the PCB to the metal case. The clearance is only around 1.2mm, provided by the pot mounting nuts, so make sure all the component leads are cut very short. This is easy using flush side-cutters on boards with plated through-holes

Using it

Fuzz boxes are most commonly used with an electric guitar or

Photo 14: the fuzz box being used with a bass synth. In this case, an open-source kit for an improved TB-303 from Adafruit, the x0xb0x (pronounced "zocks-box"). It's well worth building one: https://www.ladyada.net/make/x0xb0x/

bass. However, these days they are also used to add a bit of dynamic grunge to what may be sterile synth bass lines.

Using a fuzz box with the Roland TB-303 Bassline synthesiser, shown in Photo 14, is now a techno staple, with the fuzz adding a harmonic boost to the TB-303's accent, giving a wonderful 'bark' to the sound.

A bit of fuzz on distant filtered vocals can also work wonders to give a 'military comms' sound, such as the one in Kate Bush's nuclear holocaust song *Breathing*, coldly calculating of size of a nuclear explosion.

Distortion effects still work better in analog rather than digital, since the harmonics generated cause aliasing products that fold down into the audio.

A very high sampling rate is thus needed to perform similar effects digitally, which consumes valuable processing power that most studio engineers would rather apply to other plug-ins. Besides, this fuzz box is very portable! **PE**

Parts List – Transmanium fuzz box extras

1 assembled main PCB, 85×80mm (AO-NOV25-01) – details in last month's article	
1 jack PCB, 72.5×26mm (AO-DEC25-01)	
1 Hammond 1590 DD or Tayda A-5141 case	**[Rapid 30-4687]**
3 pointer knobs, Tayda "Marshall" type A-1838 or Action Hardware KPM-1674	**[Rapid 32-0175]**
3 80 × 3mm cable ties	**[Rapid 04-0630]**
1 CPC CBBR7741 screw-mount cable tie	
1 cable tie mounting clip/eyelet	**[Rapid 04-0600]**
1 PP3 battery snap/holder	**[Rapid 18-0093]**
1 Comf BH980 battery mounting spring clip	**[Rapid 18-3480]**
1 solderless LED socket with leads	**[Rapid 55-0167]**
1 Kingbright L-7104SEC diffused orange 3mm high-efficiency LED	**[Rapid 72-8972 Tayda A-264]**
1 TruConnect 2.1mm inner diameter, 5.5mm outer diameter DC barrel power socket	**[Rapid 20-0970, Tayda A-4118]**
2 stereo 6.3mm PCB-mount jack sockets	**[Tayda A-5079]**
1 1m length of 7/0.2 equipment wire	
1 200mm length of 3-5mm screened cable	
3 pot shaft adaptors	**[Tayda A-7023]**
3 Vishay ACCTR511-3002 ⅜-inch pot nuts	**[Mouser 594-511-3002]**
1 hex nut for Alpha potentiometers	**[Tayda A-5037]**
6 12.5mm rubber hole plugs	**[RS 605-661]**
5 M3 × 12mm countersunk head machine screws and nuts	
Pot grub screws (if needed)	**[Tayda A-7804 (M3), Tayda A-7805 (M4)]**

All parts available individually from the AOShop.

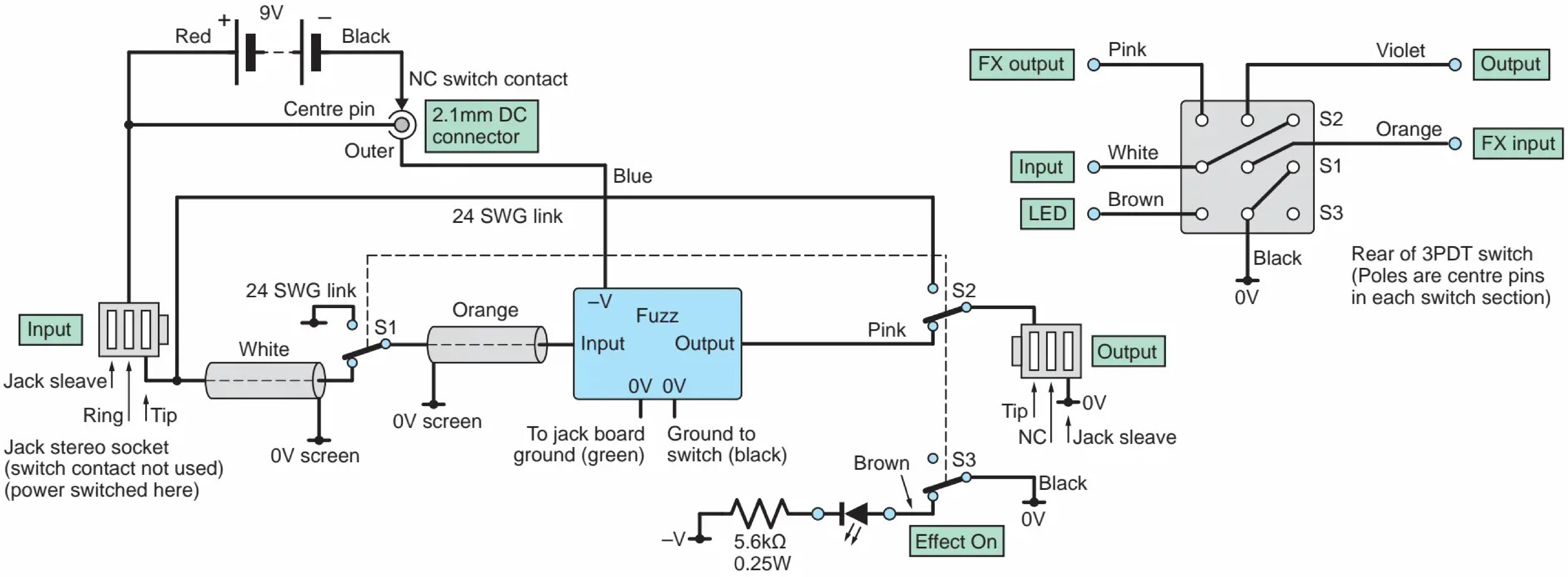
9V
Red
+
–
Black
NC switch contact
Centre pin
2.1mm DC connector
Outer
Blue
24 SWG link
24 SWG link
Input
Jack sleave
Ring
Tip
Jack stereo socket
(switch contact not used)
(power switched here)
White
0V screen
S1
Orange
0V screen
–V
Fuzz
Input
Output
0V 0V
To jack board ground (green)
Ground to switch (black)
Pink
S2
Brown
S3
–V
5.6kΩ
0.25W
Effect On
Black
0V
Output
Tip
NC
0V
Jack sleave
FX output
Pink
Violet
Output
S2
Orange
FX input
Input
White
S1
LED
Brown
S3
Black
0V
Rear of 3PDT switch
(Poles are centre pins
in each switch section)

Fig.3: all the connections in the fuzz box that are made off the circuit board. If you're unsure while wiring it, check this!

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Gyroscopes and giants

Techno Talk

Max the Magnificent

From origami-inspired circuits to dungeon-delving inertial sensors, join me on an adventure where ancient folding arts, cutting-edge gyroscopes and underground navigation all collide in unexpectedly entertaining ways.

Do you remember *Up Pompeii!*, the classic British TV comedy that aired in 1969 & 1970, starring Frankie Howerd as the ever-scheming Lurcio? One of the best-remembered running gags was Frankie's parody of an old vaudeville line when he said, "A funny thing happened to me on the way to Colosseum".

The reason this popped into (what I laughingly call) my mind is that something funny happened to me the other day. I was at home, ensconced in my comfy chair, happily reading my copy of the October 2025 issue of *Practical Electronics*, which had finally managed to wend its way across the pond to me here in America.

I'd just begun to peruse my *Techno Talk* column (by the time the magazine reaches me, I have usually forgotten what I wrote about). "Hey", said my wife (Gigi the Gorgeous), "there's a picture of some bone conduction headphones like yours in the magazine you're reading". I had to laugh. "These are my headphones", I replied. "I wrote this article!" I'd like to say she was impressed, but… moving on…

Origami, kirigami and circuits

Life is a funny old thing when you come to think about it. Back in the mists of time, when dinosaurs ruled the Earth and I was in high school, it never occurred to me that one of my chums would one day grow up to become a world-renowned origami grandmaster, but such is indeed the case.

My friend, whom we will call Nick (because that's his name), has written more than 100 books on the art of origami (nickorigami.com). The reason I mention this here is that I recently came across an interesting article describing how researchers are exploring traditional Japanese paper folding (origami) and paper cutting (kirigami) techniques to create kiri-origami printed circuit structures (https://pemag.au/link/ac8t).

On a somewhat related note, I also ran across a column describing how scientists have developed what they describe as a "photonic origami" method of producing tiny glass structures (https://pemag.au/link/ac8u). Apparently, this accidental discovery could revolutionise computing, cameras and sensors. Speaking of sensors…

Sophisticated sensors

I love the incredible, advanced sensors that are available to us these days. When I started out in electronics as a 12-year-old in 1969, the best I could hope to lay my hands on were items like magnetic reed switches, mercury tilt switches and light-dependent resistors (LDRs).

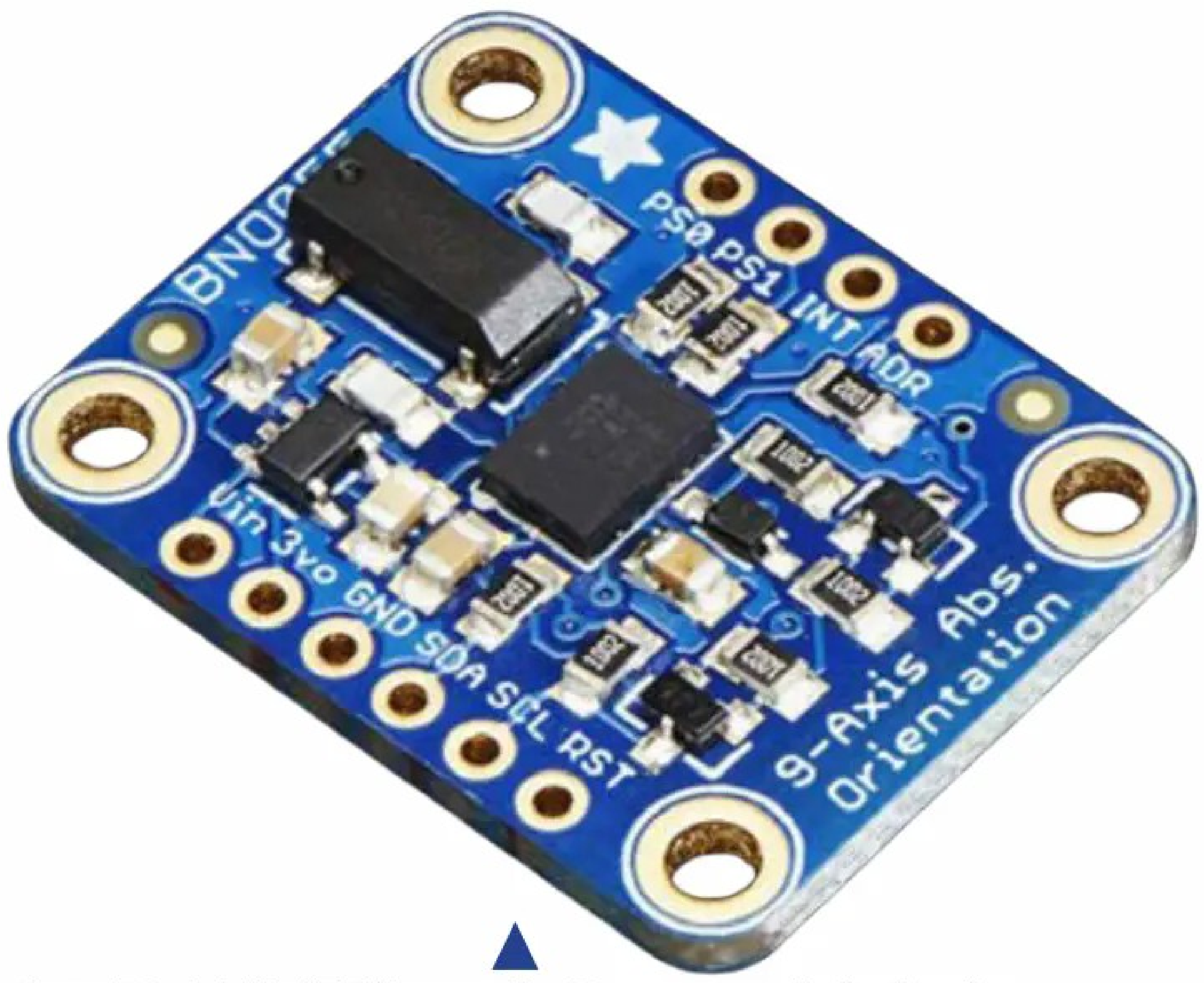
◀ Kirigami is similar to origami except that the paper may be cut and glued.

Today, you can purchase a cheap-and-cheerful breakout board (BoB) featuring a teeny-tiny chip that boasts a 32-bit microcontroller along with a micro-electromechanical system (MEMS) that provides a 3-axis accelerometer, a 3-axis gyroscope and a 3-axis magnetometer.

A 3-axis accelerometer can measure linear acceleration (change in velocity) in the X (forward-back), Y (left-right), and Z (up-down) axes. Meanwhile, a three-axis gyroscope can measure angular velocity (how fast something is rotating) around the X (roll), Y (pitch), and Z (yaw) axes. A magnetometer measures magnetic fields and can be used like a 3D compass.

Each of these devices has shortcomings when used alone: accelerometers have a noisy output and can't distinguish between tilt and linear motion, gyroscopes provide smooth short-term data but suffer from drift that accumulates over time, and magnetometers offer absolute heading but are easily disturbed by local magnetic fields.

This is where the microcontroller comes in. By applying sensor fusion algorithms, the strengths of each technology can be used to compensate for the weaknesses of the others.



▲ A 9-DoF IMU board. Source: Adafruit.

Techno Talk

Max the Magnificent

Going down…

As wonderful as the aforementioned sensors are, before using them, you need to establish a starting point from which all subsequent measurements will be made.

Suppose you were poised to enter one of the world's deepest mines, for example. You would probably begin by establishing your precise location using one or more global navigation satellite systems (GNSSs).

Examples of currently available GNSSs include GPS (developed by the USA), GLONASS (developed by Russia), Galileo (developed by the European Union), and Beidou (developed by China).

Once you'd entered the mine and transitioned into a GNSS-denied state, you might switch over to some form of inertial navigation system (INS).

This INS would employ an inertial measurement unit (IMU) to estimate your current position, velocity, and orientation without relying on external references. It's the IMU that would contain the accelerometers, gyroscopes and magnetometers.

The idea is that you could travel all the way down to the bottom of the mine, and subsequently return to the surface, while always understanding exactly where you were and in what direction you were pointing. Once again, however, this depends on knowing your starting point.

Dungeons & Dragons

Let's flip everything around a bit. Imagine that the following text is being read out in the dulcet tones of James Earl Jones, pretending to be the Dungeon Master in a game of Dungeons & Dragons (D&D): You wake up and find yourself in a huge cavern at the bottom of one of the world's deepest mines. Fortunately, the lights are on; otherwise, you really would be in a pickle.

Tunnel entrances are all around the cavern. There's also a note that says, "Take the tunnel heading north (all the others lead to certain death)".

There's an IMU next to you, but it's powered down, meaning it has lost track of its current position and orientation.

Even worse, when you look inside, you discover that some scamp has removed its magnetometer.

However, even if the magnetometer were still present, you would be foolish to rely on it because mines often contain large amounts of ferromagnetic material that distort the Earth's magnetic field and cause local anomalies, making readings unreliable.

Worst of all, you are starting to feel a little peckish. What are you going to do?

21st-century gyroscopes

Well, if you are in possession of a Boreas IMU from the Australian company Advanced Navigation (**https://pemag.au/link/ac8v**), you're in luck, because this bodacious beauty boasts three-axis cutting-edge digital fibre-optic gyroscope (DFOG) technology.

The DFOGs used in Boreas IMUs are so sensitive that they can detect the Earth's rotation, which is 15° per hour.

This means several things. First, these are the only gyroscopes on the planet that don't suffer from significant drift. Secondly, it knows which way it's pointing, even when deep underground.

This means you can quickly identify the exit tunnel in our D&D scenario and head home for a well-deserved bacon sandwich.

Recently, when I was chatting with the folks at Advanced Navigation, they told me that they had used their DFOG technology to demonstrate a breakthrough in underground navigation—delivering high-precision positioning without reliance on fixed infrastructure or GNSS.

This was demonstrated in one of the world's deepest underground mines at Pyhäjärvi, Finland, shown in the photograph below.

So, from optical origami to fibre-optic gyros, who can say which curious contraptions tomorrow will bring? As for me, I'll settle for an IMU that can guide me safely to the fridge—and back again before Gigi notices. PE

The Pyhäsalmi Mine in Pyhäjärvi, Finland. Source: Advanced Navigation.

MAX'S COOL BEANS

By Max the Magnificent

WEIRD & WONDERFUL ARDUINO PROJECTS

Part 12: storing, updating and displaying scores

Rather than wiffle-waffle, as is my wont, how about we simply plunge headfirst into the fray with gusto and abandon (always with aplomb, of course)?

You may recall that we are using 7-segment display modules as part of the retro games console we are constructing. These modules are available with two, three or four digits. I've decided to use three two-digit modules for my console.

You may opt for an alternative arrangement if you wish, but I'll assume you are working with the same setup as me for the purposes of these discussions (I'll note areas you might need to tweak to work with an alternative setup).

In our previous column, we plugged one of our two-digit modules into our breadboard. These modules feature the same 74HC595 8-bit serial-in, parallel-out (SIPO) shift registers that we've been using in our prototype experiments.

One slight 'gotcha' we encountered is that I had assumed the 7-segment displays on our modules would be of the common cathode (CC) variety, in which case 0s would turn segments **off** while 1s would turn them **on**.

I should have known better than to assume anything, because we quickly discovered that my modules boast common-anode (CA) displays. This means that it's the 0s that turn our segments **on**, while the 1s turn them **off**.

Although this issue can be easily addressed in software, it's unfortunate for several reasons. For example, our 8-segment bar graph display is currently configured in a pseudo-CC mode, with the shift register driving its anode terminals and all of its cathodes connected (via pull-down resistors) to one of the 0V rails.

As a result, when we loaded a value into our shift registers—the one on the breadboard and two more on the module—we observed the results depicted in Fig.1(a).

The term "Cycle" in this context refers to us loading values into our shift registers in 8-bit 'chunks'. Also, the digits on the modules are connected from left-to-right. Thus, the left-hand digit on the module corresponds to the value on the 8-segment display, with the right-hand digit lagging by one cycle.

You can see a full layout for the existing breadboard setup in the file named **CB-dec25-brd-01.pdf**. As usual, all files mentioned in this column are available as part of the December 2025 download package from the PE website at https://pemag.au/link/ac8p

Don't worry, be happy

I know that it shouldn't bother me that an active light-emitting diode (LED) segment on the 8-segment display corresponds to an inactive segment on the

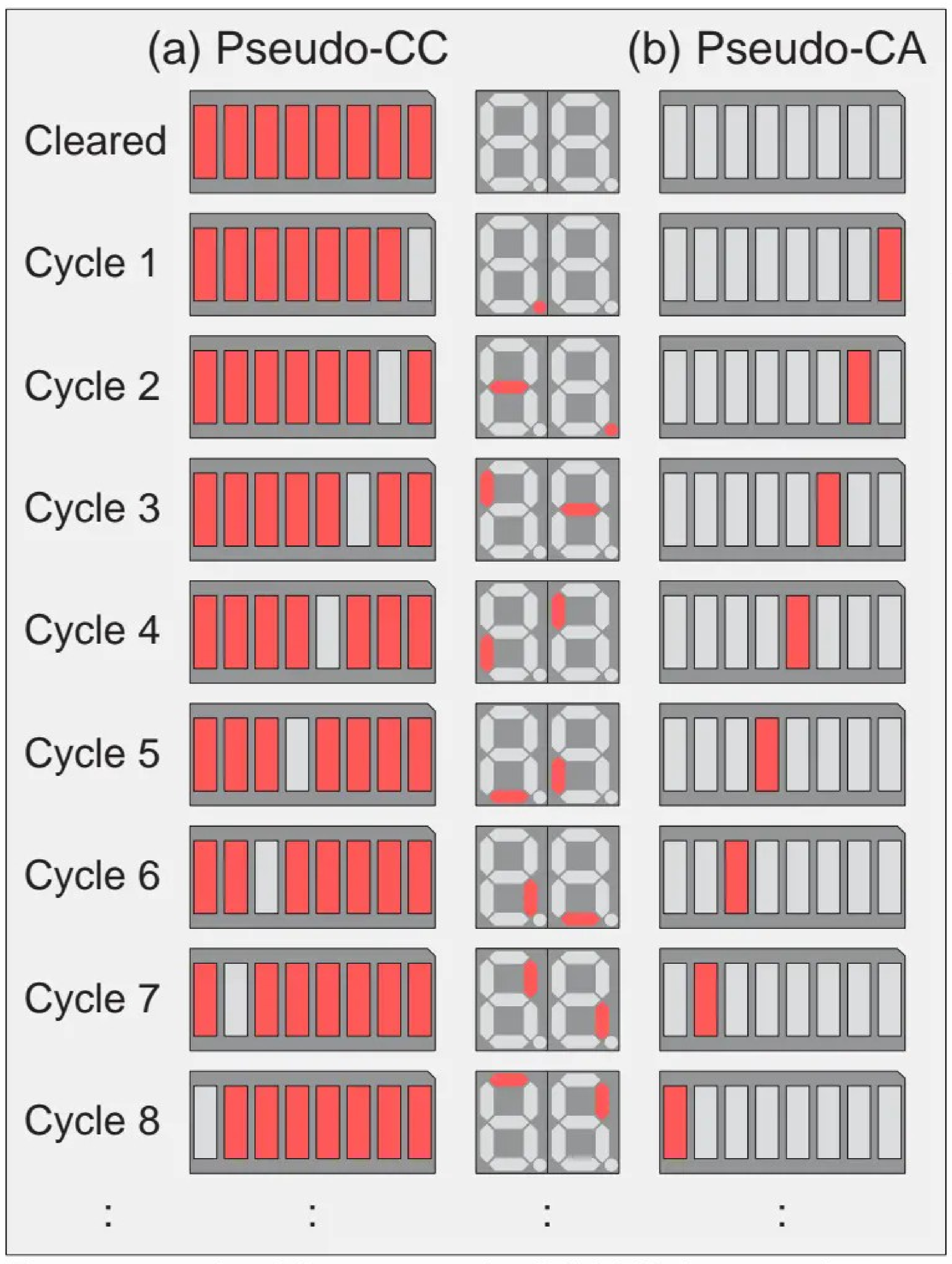


Fig.1: pseudo-CC vs pseudo-CA LED bargraphs.

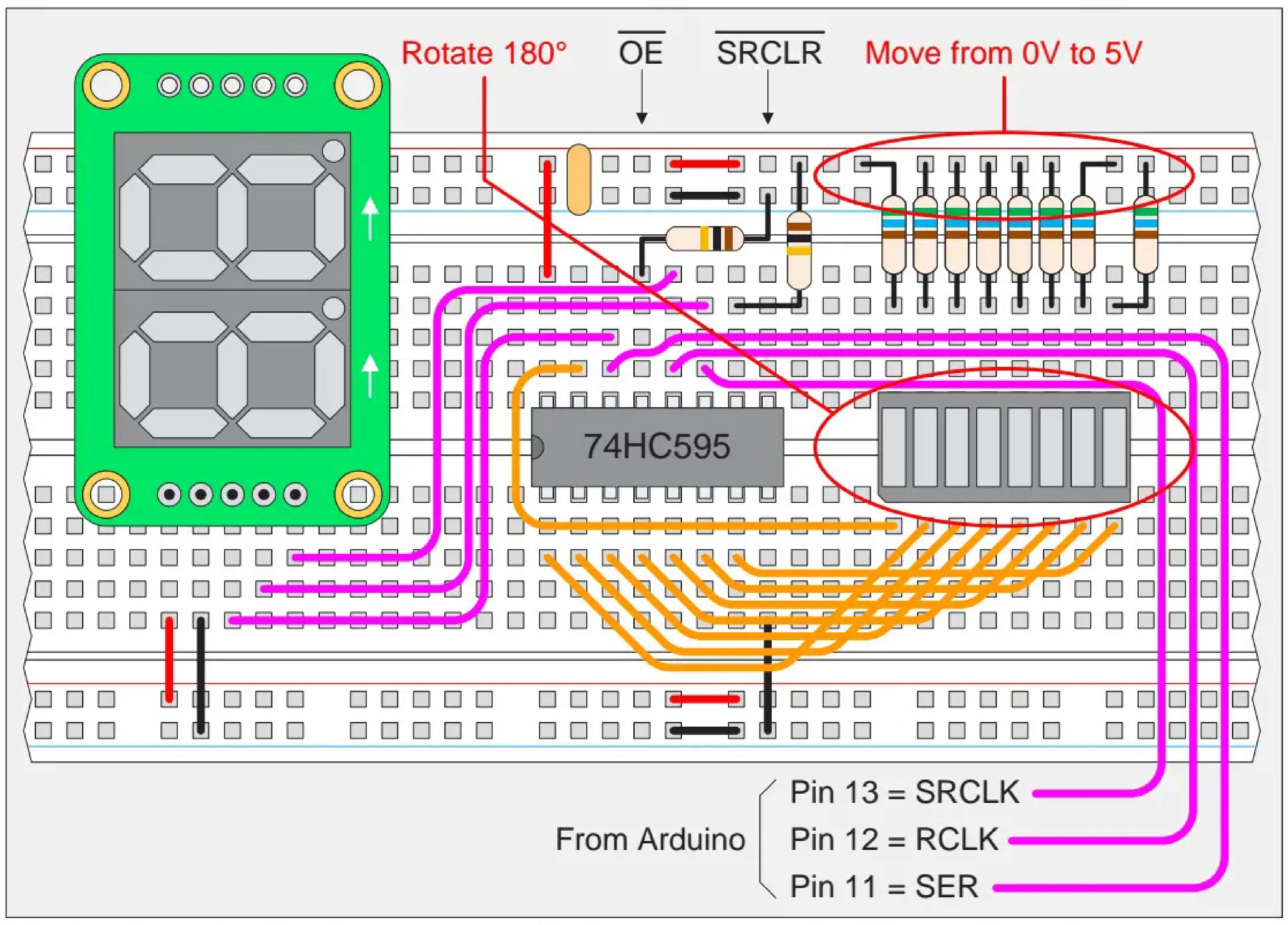


Fig.2: changing the 8-segment display to a pseudo-CA configuration.

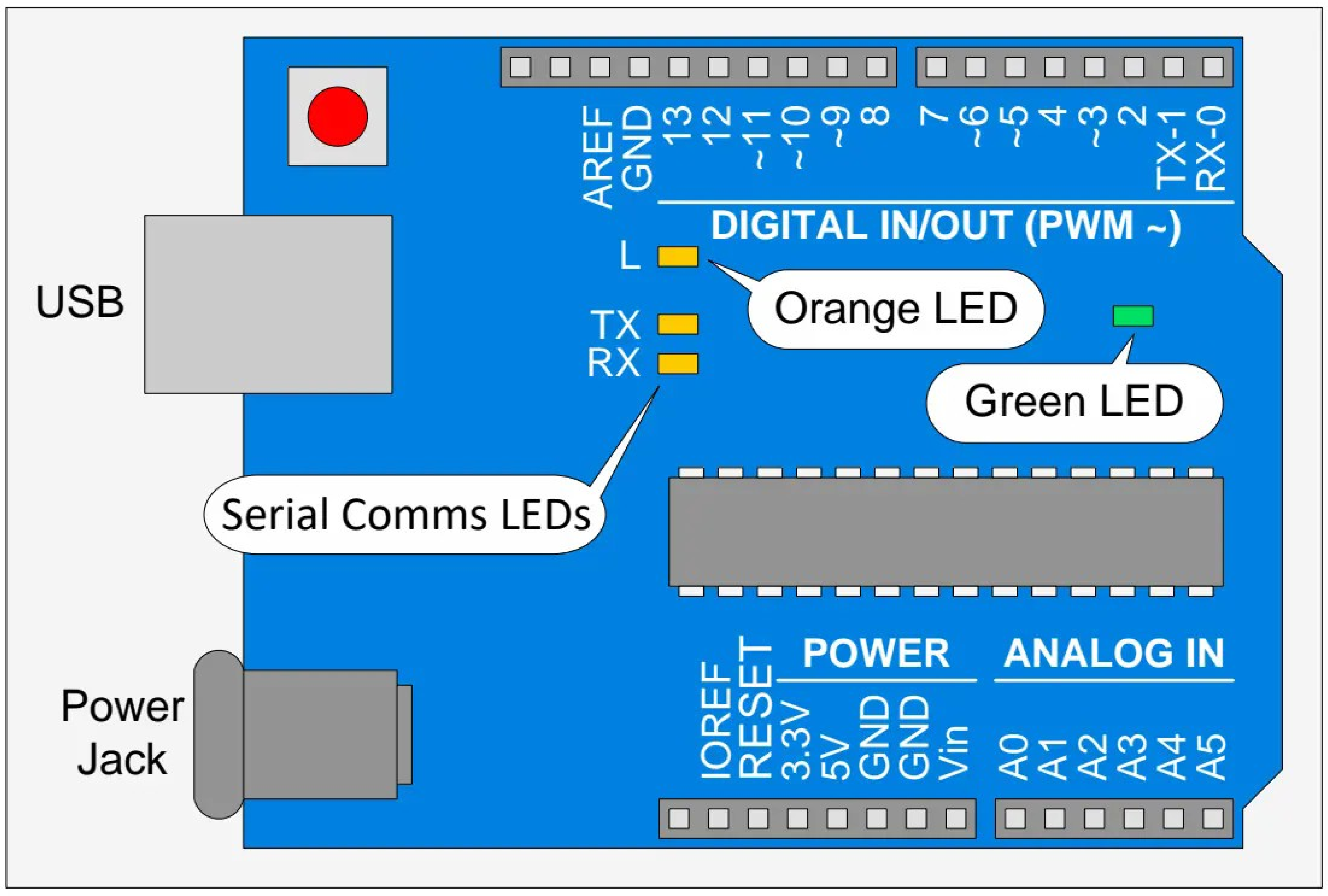


Fig.3: the four onboard LEDs on an Arduino Uno.

7-segment display, and vice versa… but it does. Things like this tend to niggle away at (what remains of) my mind. I have therefore rewired my breadboard so that my 8-segment display is now configured in a pseudo-CA mode.

All that is required is to rotate the display by 180°, which means the shift register is now driving its cathode terminals. This also brings the display's anode terminals into contact with the resistors. Furthermore, we now connect the other ends of the resistors to the +5V rail (thereby modifying their function from acting as pull-down resistors to playing the role of pull-up devices).

This new setup is illustrated in Fig.2 (the full layout is available in the file named **CB-dec25-brd-02.pdf**).

Now, when we rerun our latest and greatest program from last month (which I've made available in a file named **CB-dec25-code-01.txt**), we see the results depicted in Fig.1(b). I don't know about you, but this makes me feel much happier inside.

Ooh, flashy!

The Arduino Uno has four onboard LEDs (Fig.3). Thus far, we've only mentioned two of them: the power LED (usually green), which is lit whenever power is applied to the board, and the orange (or yellow) LED with the **L** annotation that's connected via an onboard resistor to digital input/output (I/O) pin 13. As we've seen, we can control the **L** LED with our programs.

The first thing we should do when connecting the USB cable, or otherwise powering the board, is to check that the green LED is glowing. If not, there's no use proceeding further until this problem is resolved.

If you look closely at the two remaining orange (or yellow) LEDs that are mounted close to the USB connector, you'll see that they have **RX** and **TX** annotations next to them. Have you noticed that when you upload a program (sketch) to your Arduino Uno, the **L** LED flashes first, followed by the **RX** and **TX** LEDs flashing repeatedly? What's happening here?

Well, the bootloader (see the panel on page 44) first flashes the **L** LED as a simple status indicator to show that it's alive and functioning. It then flashes the **RX** and **TX** LEDs to indicate that serial communication is taking place between your computer and the Arduino.

Your main computer may be referred to as the 'host' because it initiates the communication and controls the process. The use of this in computing (and networking) is derived from the social sense of host; that is, the person who invites guests, provides resources (food, drink, space), and coordinates the event.

The device with which the host is communicating (the Arduino in this case) may be referred to as the 'target'. Just like what your dinner host calls you (I must confess that my dinner parties are a bit odd).

Anyway, the **RX** LED blinks when the Arduino is *receiving* data *from* your computer, while the **TX** LED blinks when the Arduino is *transmitting* data *to* your computer. During an upload, both of these LEDs will usually flash rapidly.

Assuming there aren't any errors in our code, when we click the "Upload" icon on the Arduino's IDE, it will report a series of four messages: **Compiling → Done Compiling → Loading → Done Loading**, after which the program will automatically run. I don't know about you, but personally, I like watching the **L**, **TX**, and **RX** LEDs flash while I'm uploading a program.

Even though the IDE prints messages like "Compiling" and "Uploading", the flashing **L**, **RX** and **TX** LEDs are real, physical indicators. They provide tangible reassurance that something is really happening.

Flashing LEDs are crude but surprisingly effective diagnostic tools. If the **L**, **RX** and **TX** remain dark when we expect activity, we know something is wrong before the IDE times out. There's also an aspect of 'trust but verify'. The text messages in the IDE provide one layer of confirmation, while LEDs offer a second, independent channel.

Why am I waffling on about all this here? Well, returning to Fig.2, we observe that the **SER**, **RCLK** and **SRCLK** signals to the shift registers are driven by Arduino pins 11, 12, and 13, respectively. The reason I used pin 13 to drive **SRCLK** is that this signal will toggle the most when we are loading new values into our 7-segment display modules.

Since pin 13 is also connected to the **L** LED on the Arduino, it provides a palpable indication that something is happening.

In reality, things happen so fast in our current code that we can't perceive anything. I added a new definition called **STILL_KICKING**, assigning a value of 1, which will be used to add a 1ms (one millisecond) delay to each pulse of the **SRCLK**. I then added this delay to our **LoadSR()** ("load shift register") function, as seen on line 75 in Listing 2(a).

Feel free to download the new version of our program (the file named **CB-dec25-code-02.txt**). Remember, we're using the convention that the listing number (2 in this case) corresponds to the numerical part of the matching code file name ("02" here).

We might change the way we've implemented our blinkenlight in the future (or remove it completely) but I'm really enjoying its company for the moment.

Staying in character

Thus far, we've illuminated only one segment at a time. The next step will be to light up multiple segments

```
68  // Load 8-bit value into the shift register
69  void LoadSR (uint8_t thisByte)
70  {
71    for (int iSeg = 0; iSeg < NUM_SEGS; iSeg++)
72    {
73      digitalWrite(PinSER, (thisByte & 1));
74      digitalWrite(PinSRCLK, SR_CTRL_ACTIVE);
75      delay(STILL_KICKING);
76      digitalWrite(PinSRCLK, SR_CTRL_INACTIVE);
77      thisByte >>= 1;
78    }
79  }
```

Listing 2(a): adding a "still kicking" delay.

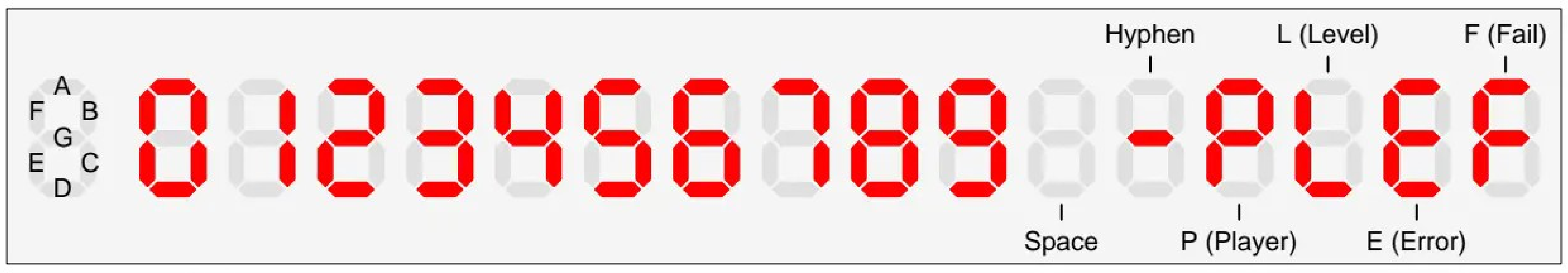


Fig.4: our initial 7-segment character set.

simultaneously to form characters, but which characters do we want to represent?

For no other reason than 'just because', I've decided to limit myself to only 16 different characters. Apart from anything else, we can reference 16 characters using a 4-bit field. Having said this, it's not that we are short of memory (at least, not yet), and there's no reason why we couldn't extend our character set in the future if we so desire.

Since we are building a retro games console, we know that we will want to display scores, meaning we need to be able to represent the ten numerical characters in the range 0 to 9. This leaves us with six empty 'slots' in our 16-character set.

Off the top of my head, I think space (blank) and hyphen (minus) characters might come in handy. Also, we could use "P" for "Player" (in multi-player games) and "L" for Level (in multi-level games), along with "E" and "F" for "Error" and "Fail", respectively (I'm sure we will find a use for these at some stage in the proceedings).

Therefore, the character set we will implement at this time is as depicted in Fig.4. Note that we currently have no plans to employ decimal point segments, but that could change later if we need them.

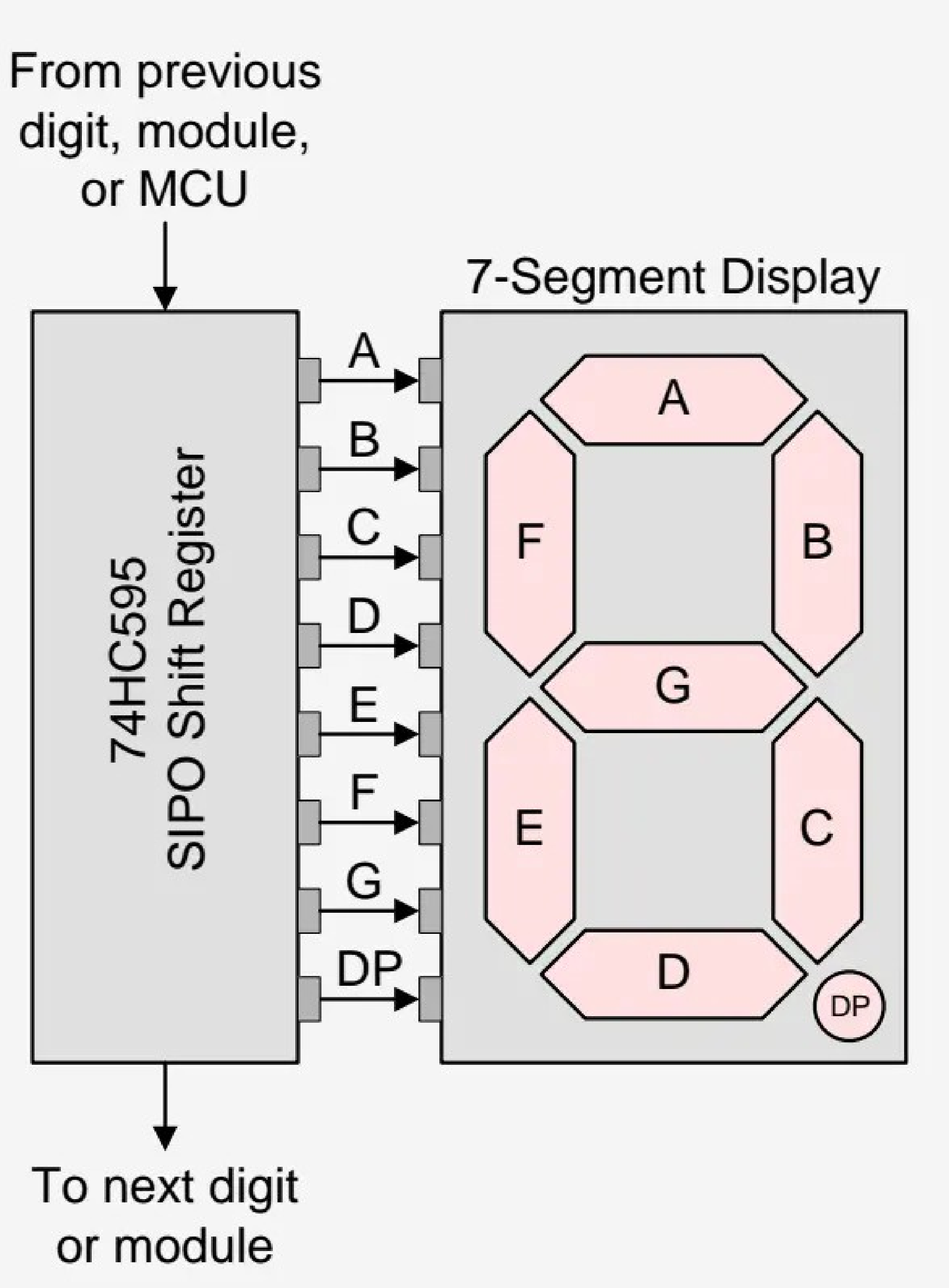
Talented tables

From last month's experiments, we know that when we load an 8-bit byte into the shift register driving a 7-segment display on our modules, the first bit, which we are thinking of as being the least-significant bit (LSB), ends up driving the **DP** ("decimal point") segment. Similarly, the last bit, which we are thinking of as being the most-significant bit (MSB), ends up driving the **A** segment.

Armed with this knowledge, we can represent everything in tabular form, as in Fig.5. We've shown the **DP** column as being greyed out, reflecting the fact that we have no plans to use this segment at this time (we still need to load a 0 into the shift register bit associated with this segment).

Actually, since the 7-segment displays in my modules are of the common anode (CA) flavour, a 0 in the shift register will turn the corresponding segment on, while a 1 will turn it off. So we actually need to load a 1 into the shift register bit to keep the DP off. But I still think of it as 0 = off because that's more intuitive to me.

As we discussed last month, we can represent things this way in our programs, then invert the values at the last moment just before we feed them to the shift register. If we ever find ourselves driving common cathode (CC) displays, all we need to do is to omit the final inversion.


From previous digit, module, or MCU
7-Segment Display
74HC595 SIPO Shift Register
A
B
C
D
E
F
G
DP
A
B
C
D
E
F
G
DP
To next digit or module

Index	To 7-Segment A B C D E F G DP	Display
0 (0x0)	1 1 1 1 1 1 0 0	0
1 (0x1)	0 1 1 0 0 0 0 0	1
2 (0x2)	1 1 0 1 1 0 1 0	2
3 (0x3)	1 1 1 1 0 0 1 0	3
4 (0x4)	0 1 1 0 0 1 1 0	4
5 (0x5)	1 0 1 1 0 1 1 0	5
6 (0x6)	1 0 1 1 1 1 1 0	6
7 (0x7)	1 1 1 0 0 0 0 0	7
8 (0x8)	1 1 1 1 1 1 1 0	8
9 (0x9)	1 1 1 1 0 1 1 0	9
10 (0xA)	0 0 0 0 0 0 0 0	Space
11 (0xB)	0 0 0 0 0 0 1 0	Hyphen
12 (0xC)	1 1 0 0 1 1 1 0	P
13 (0xD)	0 0 0 1 1 1 0 0	L
14 (0xE)	1 0 0 1 1 1 1 0	E
15 (0xF)	1 0 0 0 1 1 1 0	F

Fig.5: mapping segments to characters.

The next consideration involves how we will load these bit patterns into our shift register chain. Take the character "5", for example. From Fig.5, we know that the bit pattern associated with this character is 10110110. Remembering that we are inverting everything to accommodate my CA displays, the bit pattern we will need to load into the shift register is 01001001.

Additionally, we know that we will start loading from the least significant (right hand) bit. Based on this, we could do things explicitly as follows:

- Set SER to 1, pulse SRCLK once
- Set SER to 0, pulse SRCLK twice
- Set SER to 1, pulse SRCLK once
- Set SER to 0, pulse SRCLK twice
- Set SER to 1, pulse SRCLK once
- Set SER to 0, pulse SRCLK once

I don't know about you, but having to code this sort of thing by hand for each digit in our display would be horribly inefficient and make my brain leak out of my ears. Instead, we will let our Arduino handle all the heavy lifting for us.

Updating our program

In a moment, we will update our program to present honest-to-goodness characters on our display modules. First, however, we must decide which of the displays will form our least significant digit (LSD) and which will act as our most significant digit (MSD).

This is another case that is somewhat arbitrary. For example, since data is loaded from left-to-right into our two-digit modules, some people may prefer to think of the leftmost digit as being the LSD. For myself, given a choice, I always prefer to view the LSD as being the one on the right and the MSD as being the one on the left.

This means that Fig.6 depicts the way I view the two-digit module I'm currently working with. As usual, regardless of which approach we decide to take, we can always craft our code to make things function the way we want.

Keep this in mind as we consider our new program (in the file named

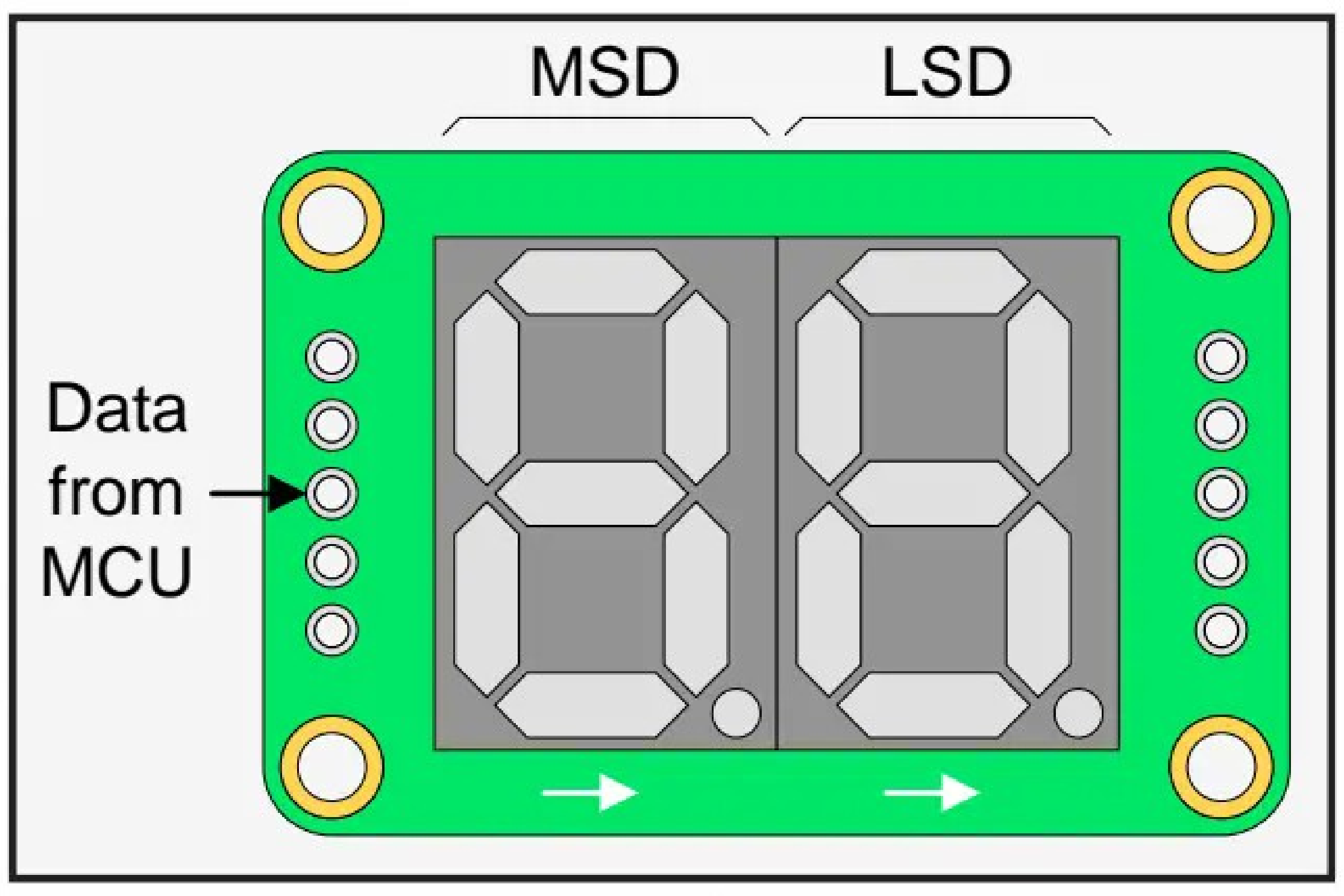


Fig.6: a 1-module (2-digit) display.

```
3    // Display Stuff
4    #define NUM_SEGS    8 // Segments per display
5    #define NUM_DIGS    2 // Number of digits
6    #define ALL_SEGS (NUM_SEGS * NUM_DIGS)
7    #define NUM_CHARS  16 // Characters
8
9    #define SPACE      10
10   #define HYPHEN     11
11   #define P_CHAR     12
12   #define L_CHAR     13
13   #define E_CHAR     14
14   #define F_CHAR     15
15
16   // Shift Register (SR) Stuff
17   #define SR_CTRL_INACTIVE   LOW
18   #define SR_CTRL_ACTIVE    HIGH
19   #define SR_DATA_INACTIVE  HIGH
20   #define SR_DATA_ACTIVE     LOW
21
22   // Extras
23   #define UPDATE_RATE       1000
24   #define STILL_KICKING        1
```

Listing 3(a): our updated definitions.

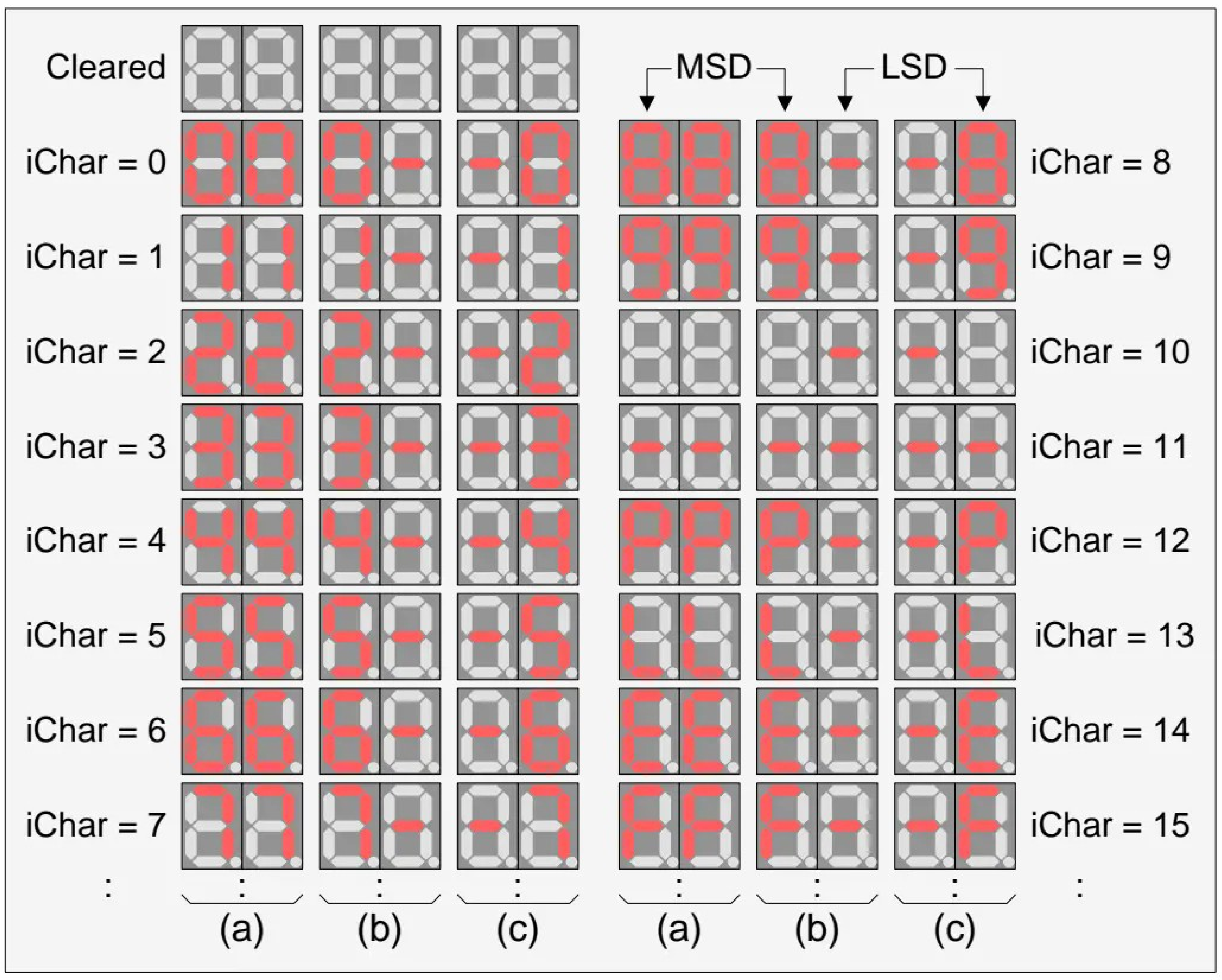


Fig.7: the results of our test code – success!

```
26   // Digits to be displayed
27   uint8_t DisplayBuffer[NUM_DIGS];
28
29   // Bit patterns for characters
30   uint8_t CharSegMap[NUM_CHARS] =
31   {
32   //  ABCDEFGDP
33     0b11111100, //  0 (0x0) 0
34     0b01100000, //  1 (0x1) 1
35     0b11011010, //  2 (0x2) 2
36     0b11110010, //  3 (0x3) 3
37     0b01100110, //  4 (0x4) 4
38     0b10110110, //  5 (0x5) 5
39     0b10111110, //  6 (0x6) 6
40     0b11100000, //  7 (0x7) 7
41     0b11111110, //  8 (0x8) 8
42     0b11110110, //  9 (0x9) 9
43
44     0b00000000, // 10 (0xA) Space
45     0b00000010, // 11 (0xB) Hyphen
46     0b11001110, // 12 (0xC) P
47     0b00011100, // 13 (0xD) L
48     0b10011110, // 14 (0xE) E
49     0b10001110  // 15 (0xF) F
50   };
```

Listing 3(b): additional global variables.

CB-dec25-code-03.txt). This is based on our previous offering. You may be surprised by the modest modifications required (or not, as the case might be). Let's start with the definitions as shown in Listing 3(a).

Many of these remain unchanged. On line 7, we've added **NUM_CHARS** and assigned it a value of 16. This reflects the number of different characters we wish to depict, as shown in Fig.4. And on lines 9 through 14, we create definitions for the index values associated with our special characters (I'm sure these definitions will come in handy).

Next, we add two new global variables. The first, **DisplayBuffer[]**, is an array of unsigned 8-bit values. Each element of the array represents one of the digits on our display. This will act as a buffer to store the bit patterns we wish to upload into our shift register chain.

The second, **CharSegMap[]**, is also an array of unsigned 8-bit values. In this case, we assign it with the bit patterns associated with the various characters we may wish to display, as depicted in Fig.5.

The third and final area of change can be seen in our spiffy new **loop()** function, as seen in Listing 3(c). On line 78, we declare a **for()** loop that will cycle through each of our characters.

On lines 80 and 81, we load both elements of our display buffer with the bit pattern associated with the character currently pointed to by our **iChar** variable. Next, on lines 83 through 86, we call our **LoadSR()** function to copy the two 8-bit patterns stored in the **DisplayBuffer[]** array into our shift register chain.

On line 87, we call our **LoadOR()** ("load output register") function to copy the contents of the shift registers into their corresponding output registers. Finally, on line 88, we pause to cast our orbs over the characters as they appear on the display.

Observe that line 85 is where we use the **~** (bitwise NOT) operator to invert all the bits (exchanging 0s for 1s and 1s for 0s) as we pass them into the **LoadSR()** function. If your display modules feature common cathode devices, you need only delete this single-character operator for everything to work as desired. The results should be as shown in Fig.7(a).

Are we sure?

Since we are currently displaying the same character on both digits, there's always a chance that we've messed things up with respect to loading our LSD first and our MSD last (stranger things have happened).

Let's check. Modify line 80 in Listing 3(c) to change the value being assigned to **DisplayBuffer[0]** (which we intend to represent our LSD) from **CharSegMap[iChar]** to **CharSegMap[HYPHEN]**. Now, the results should be as shown in Fig.7(b).

Return line 80 to its original state, then modify line 81 to change the value being assigned to **DisplayBuffer[1]** (which we intend to represent our MSD) to **CharSegMap[HYPHEN]**. This should cause the results to appear as shown in Fig.7(c).

I love it when a plan comes together.

Keeping score

Let's move on to something a little more exciting. Depending on the game we are playing, one of the things we will typically want to do is store, manipulate (eg, increment/decrement) and display the player's current score.

To do this, let's create some new functions: one to update the current score while ensuring that it keeps

```
75   // Do this over and over again
76   void loop ()
77   {
78     for (int iChar = 0; iChar < NUM_CHARS; iChar++)
79     {
80       DisplayBuffer[0] = CharSegMap[iChar];
81       DisplayBuffer[1] = CharSegMap[iChar];
82
83       for (int iDigs = 0; iDigs < NUM_DIGS; iDigs++)
84       {
85         LoadSR(~DisplayBuffer[iDigs]);
86       }
87       LoadOR();
88       delay(UPDATE_RATE);
89     }
90   }
```

Listing 3(c): our spiffy new loop() function.

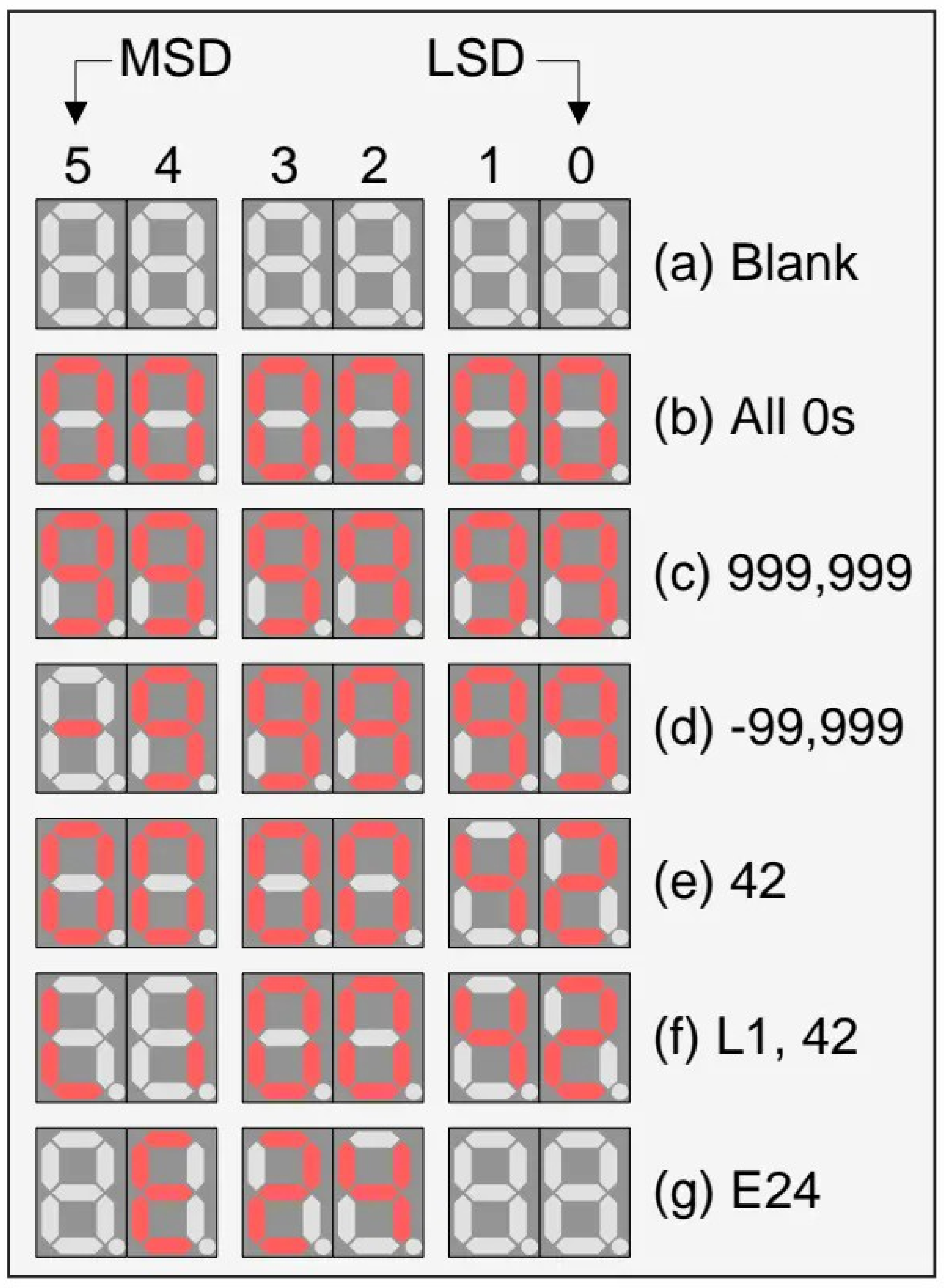


Fig.8: various alignments and widths.

within the specified bounds, one to load the score into the display buffer, and one to copy the contents of the display buffer to the display itself.

Different games will have different requirements. Rather than having to revamp our functions for each new game, it will prove to be much more efficient to develop versatile, general-purpose functions that can satisfy the needs of multiple games.

Let's start by considering that classic '70s and '80s games mostly avoided negative scores, 'clamping' the lower bound of the score to 0. They did this to keep things simple.

However, there are cases where negative scores may be appropriate, including penalties/infractions (deducting points for mistakes, which can result in totals dipping below 0) and using time/energy as a score (overtime or damage could result in a negative tally). Thus, as a starting point, let's assume that we need to be able to handle both positive and negative score values.

Remembering that we are eventually going to have three two-digit displays (although you may opt for fewer or more), the worst-case scenario using all the displays is that we want to be able to store and display values between -99,999 ("-9 99 99") to 999,999 ("99 99 99").

Are we aligned?

Since we will have six digits, we might be tempted to make two assumptions. The first is that we will always use all the digits to present our numbers (scores). The second assumption is that we will always wish to present our numbers right-aligned to our LSD. As we shall see, however, neither of these assumptions holds water.

If we were to display six-digit values, some examples could be as shown in Fig.8(b) through Fig.8(e). In the case of Fig.8(e) in particular, observe that I've made the decision to always display leading zeros (we aren't obliged to do things this way, but it does make life easier).

However, a little thought reveals that there are other possibilities. Suppose we have a game with nine levels numbered 1 to 9, for example. Further, suppose that we are currently on level 1. In this case, we might wish to use the two leftmost digits to display "L1", leaving the four remaining digits to present the four-digit score.

Assuming a current score of 42, this case is illustrated in Fig.8(f). This means we need the ability to display a 1-digit number ranging from 1 to 9 in the digit 4 position to reflect the current level, while also displaying a 4-digit number right-aligned on the LSD to display our 4-digit score.

As another example, suppose we want to display an "Error" value between 0 and 99. In this case, as illustrated in Fig.8(g), we might decide to present an 'E' character in the digit 4 position and use digits 2 and 3 to represent our two-digit error code.

We don't wish to limit ourselves to just these cases because we haven't really thought things through. That is, we don't know what games we will be creating, and we don't know how we might wish to display our numbers in the future.

Thus, the best thing to do is to say that we wish to have the ability to display numbers with 'widths' ranging from one to **n** digits (where **n** is 6 in my case), and we want to be able to specify the location of each number's least significant digit on the display. This is going to be far easier than you might imagine.

Updating a score

When it comes to storing scores (or any other numbers we might wish to present on our displays), the most straightforward approach is to simply use signed 32-bit integers. This will allow us to employ the Arduino's regular mathematical operators (addition and subtraction, plus possibly multiplication, division, and modulo).

Why do we need 32 bits? Well, an 8-bit field can contain $2^8 = 256$ patterns of 0s and 1s, which can represent signed numbers in the range -128 to +127. This isn't sufficient to accommodate our worst-case score values, which range from -99,999 to +999,999.

Similarly, a 16-bit field can be used to represent $2^{16} = 65,536$ patterns of 0s and 1s, which can be used to represent signed numbers in the range -32,768 to +32,767. Again, this isn't sufficient to accommodate our worst-case score values. This is why we must turn to 32-bit fields, which can be used to represent signed numbers in the range -2,147,483,648 to +2,147,483,647.

As you will see if you peruse the latest incarnation of our program (the file named **CB-dec25-code-04.txt**), we've added two definitions, **MIN_SCORE** and **MAX_SCORE**, to set the bounds of our score.

Since we currently have only two digits (we'll add the remaining four digits next month), we'll commence by considering only positive scores in the range 0 to 99, so we've assigned values of 0 and 99 to **MIN_SCORE** and **MAX_SCORE**, respectively.

Additionally, we've added two more definitions: **ZERO** and **MINUS**, to which we've assigned values of 0 and 11, respectively. Observe that **MINUS** has the same value as our existing **HYPHEN** definition. We will use these as pointers into our **CharSegMap[]** array to make it easier to see what's going on in our code.

When it comes to storing and manipulating scores, we've declared three 32-bit integers: **P1Score** ("player one's score"), **P2Score** ("player two's score") and **DeltaVal** (the value to be added to or subtracted from the current score). In the **setup()** function, we assign these variables the values of **MIN_SCORE**, **MIN_SCORE** and 1, respectively.

Now let's look at the **UpdateScore()** function, as illustrated in Listing 4(a).

The reason all of our previous function definitions have begun with the **void** keyword is that we haven't required them to return a value. In this case, however, we want our function to return a 32-bit integer, which is why we precede the name of the function with **int32_t** on line 101. This tells the compiler the data type of the value that our function is to return.

```
100 // Modify a score with a delta (change) value
101 int32_t UpdateScore (int32_t score, int32_t delta, int32_t minv, int32_t maxv)
102 {
103     int32_t tmpv = score + delta;
104     if (tmpv < minv) tmpv = minv;
105     if (tmpv > maxv) tmpv = maxv;
106     return tmpv;
107 }
```

Listing 4(a): updating the score while keeping it in bounds.

This function accepts four parameters: the current **score** we wish to use as a starting point, the **delta** (change) value we wish to apply, the **min** ("floor") value we will allow, and the **max** ("cap") value we will support.

One interesting point to note is that we aren't limiting ourselves to simply incrementing or decrementing the existing score by +1 or -1, respectively. The **delta** value can be any positive or negative integer.

The rest of this function should be self-explanatory. The only thing we haven't seen before is the **return** keyword, which is used to return the designated value to the point from which the function was called (we will see this in action when we call this function from within our **loop()** function).

Loading the display buffer

Next, we have the **NumberToBuffer()** function, as illustrated in Listing 4(b). The reason we called this "number to buffer" rather than "score to buffer" is that, as we previously discussed, we may wish to display numbers other than scores (eg, level and error values).

Actually, this can be a little tricky to wrap one's brain around, so perhaps one more diagram will help (Fig.9). Let's take things in reverse order. Our end goal is to present characters on our 7-segment displays.

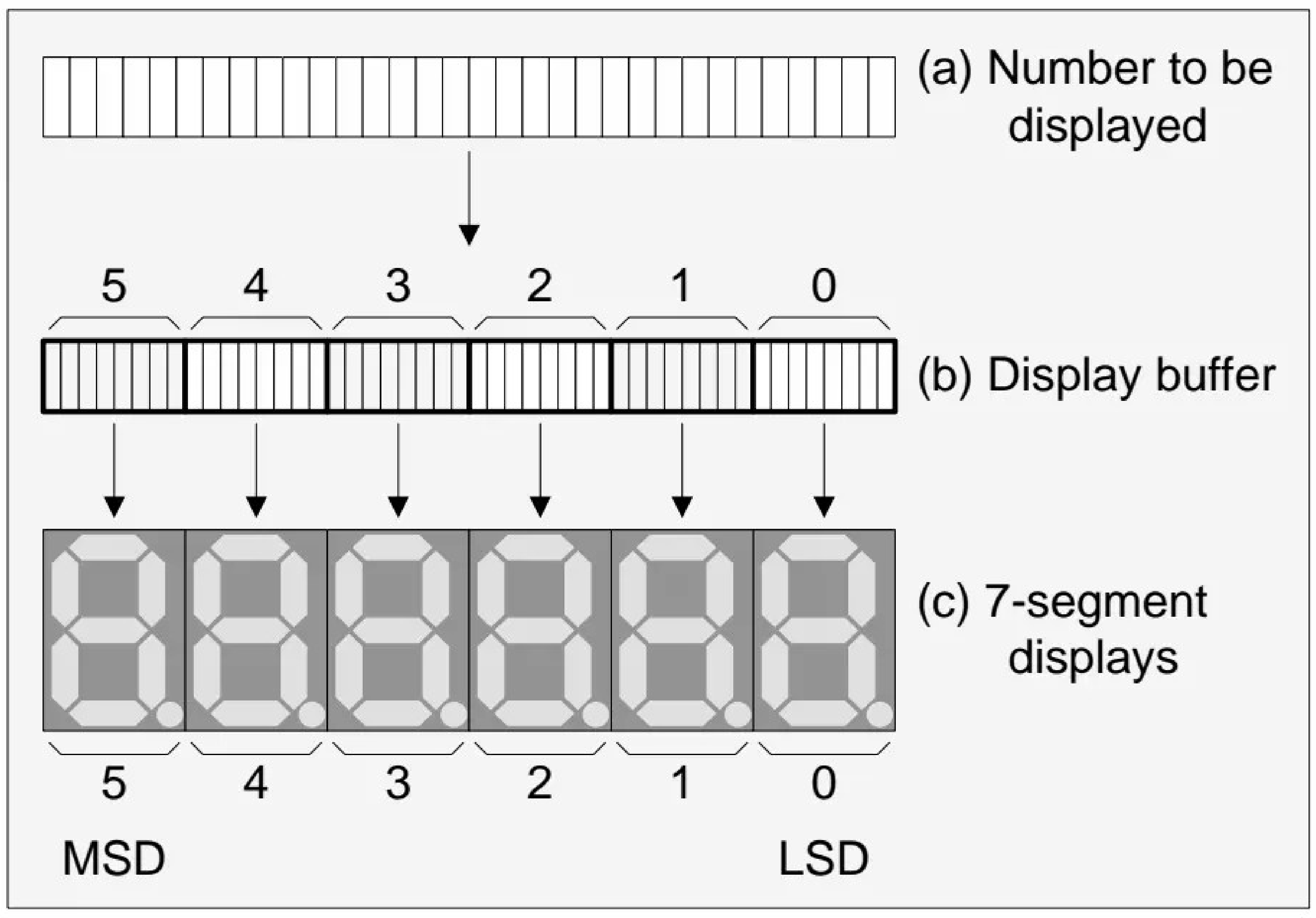


Fig.9: mapping a number onto the display.

We are assuming six such displays for the purposes of these discussions, as depicted in Fig.9(c). Each of these displays has an associated shift register, which we've omitted from this diagram to keep things simple.

We have a display buffer (called **DisplayBuffer[]**) that's defined as a six-element array. As illustrated in Fig.9(b), each of these elements is an 8-bit byte that's used to store the bitmap corresponding to the character we wish to display. We could, if we wish, load certain elements in the display buffer with the bit patterns associated with characters like "L", "E", "F" and "-".

Last but not least, we have a number we wish to display. This will be stored as a 32-bit signed integer, as shown in Fig.9(a). The important points to remember are that this number may be positive or negative, that it can occupy anywhere from one to six digits, and that the LSD of the displayed number may be offset from the LSD of the 7-segment displays.

Returning to Listing 4(a), our function **NumberToBuffer()** accepts three parameters: **value** (the number we want to load into the buffer), **numDigs** (the number of digits we wish to display) and **offset** (the number of digits by which we wish to offset the value we display from the LSD on the right of the display).

If offset is 0, the rightmost digit of the value to be displayed will be loaded into digit 0 in our buffer and appear on digit 0 of the display; if offset is 1, then the rightmost digit of the value to be displayed will be loaded into digit 1 in our buffer and appear on digit 1 of the display; and so forth.

The first thing we do is use line 113 to check to see if we are working with a negative value. If this is the case, we know we will want to display a minus sign. In turn, this means that we will have one fewer digit to display. Thus, on line 115, we reduce the value of **numDigs** by 1.

On line 116, we load the MSD associated with this number in our buffer (which isn't necessarily the MSD of the display) with the bitmap associated with a minus sign. Then, on line 117, we convert our negative value into its positive counterpart for the purposes of displaying the little scamp.

Earlier, we decided that we wished to represent all of our numerical values with leading zeros. This is accomplished by the **for()** loop that commences on line 121. This loop loads the bitmap associated with a "0" character into all the digits in our buffer that we are using to represent this number (as specified by **numDigs**).

The clever thing here is that, if we happen to be displaying a negative number, then the fact that we've already decremented the value of **numDigs** on line 115 means we won't overwrite the minus sign we loaded into the MSD associated with our number in the display buffer.

Finally, we load the number into the display buffer. As long as the value remains greater than 0 (see the **while()** loop that commences on line 127), we perform three tasks. First, we load the display buffer element specified by **offset** with the bitmap associated with the numerical character corresponding to the remainder returned when our value is divided by 10.

Second, we divide our value by 10, thereby readying ourselves for the next pass around the loop. Finally, we increment the **offset** to point to the next element in the display buffer.

I must admit that I'm rather proud of this **NumberToBuffer()** function because I wrote it off the top of my head and it worked the first time (phew!).

Copying the buffer to the display

The only remaining function is used to copy the contents of the display buffer to the display itself. This

```
109 // Load a number into the display buffer
110 void NumberToBuffer (int32_t value, int8_t numDigs, int8_t offset)
111 {
112   // Handle a negative value
113   if (value < 0)
114   {
115     numDigs -= 1; // One less digit to display
116     DisplayBuffer[offset + numDigs] = CharSegMap[MINUS];
117     value = -value; // Make the value positive
118   }
119
120   // Preload leading zeros
121   for (int iDigs = offset; iDigs < (offset + numDigs); iDigs++)
122   {
123     DisplayBuffer[iDigs] = CharSegMap[ZERO];
124   }
125
126   // Load the value itself
127   while (value > 0)
128   {
129     DisplayBuffer[offset] = CharSegMap[value % 10];
130     value /= 10;
131     offset += 1;
132   }
133 }
```

Listing 4(b): loading a number into the display buffer.

```
// Display the contents of the buffer
void BufferToDisplay ()
{
  for (int iDigs = 0; iDigs < NUM_DIGS; iDigs++)
  {
    LoadSR(~DisplayBuffer[iDigs]);
  }
  LoadOR();
}
```

Listing 4(c): copying the buffer contents to the display.

function is shown in Listing 4(c). Previously, we performed this task as part of our **loop()** function; however, it makes more sense to have it as a function in its own right.

Remember that this function neither knows nor cares what's stored in the display buffer. It just streams the bitmaps in the display buffer into the shift register chain formed by the display modules, and "Bob's your uncle" (or aunt, depending on your family dynamic).

It's time to rock!

Finally, it's time to bring everything together by means of our spiffy new **loop()** function, as depicted in Listing 4(d). Remember that we've currently defined **MIN_SCORE** and **MAX_SCORE** to be 0 and 99, respectively. Also, we've initialised the variable **P1Score** to contain **MIN_SCORE** and the variable **DeltaVal** to contain +1.

On line 92, we load the value stored in **P1Score** into our display buffer, while also specifying that we wish to represent this number using two (2) digits with an offset of zero (0). On line 93, we stream the contents of the display buffer into the display itself. On line 94, we pause to peruse what we have wrought.

Now observe line 97. When we call our **UpdateScore()** function, we pass in the current value of **P1Score** (along with the delta, min, and max values). The clever thing is that we load the value returned from this function call back into **P1Score**.

```
88   // Do this over and over again
89   void loop ()
90   {
91     // Display the current score
92     NumberToBuffer(P1Score, 2, 0);
93     BufferToDisplay();
94     delay(UPDATE_RATE);
95
96     // Modify the score
97     P1Score = UpdateScore(P1Score, DeltaVal, MIN_SCORE, MAX_SCORE);
98   }
```

Listing 4(d): orchestrating everything via the **loop()** function.

```
88   // Do this over and over again
89   void loop ()
90   {
91     // Display the current score
92     NumberToBuffer(P1Score, 2, 0);
93     BufferToDisplay();
94     delay(UPDATE_RATE);
95
96     // Reverse direction when min or max reached
97     if ((P1Score == MIN_SCORE) || (P1Score == MAX_SCORE))
98       DeltaVal = -DeltaVal;
99
100    // Modify the score
101    P1Score = UpdateScore(P1Score, DeltaVal, MIN_SCORE, MAX_SCORE);
102   }
```

Listing 5(a): this version of the **loop()** function causes the score value to 'ping pong'.

Since we've currently set **DeltaVal** to be +1, this means we will see the 00, 01, 02, 03….. 97, 98, 99 sequence appear on our 2-digit display. Additionally, since our **UpdateScore()** function will cap the score at **MAX_SCORE**, which is currently defined as 99, we will end up seeing 99, 99, 99…

Just for the thrill of it, I've created one final (for this column) iteration of our program (in the file named **CB-dec25-code-05.txt**). All I've done is to modify the **loop()** function as shown in Listing 5(a).

On line 97, we check to see if our score has reached its minimum or maximum value. If so, on line 98, we reverse the polarity of **DeltaVal**. This means our display will repeatedly count up from 00 to 99 and then back down to 00 again.

Although we currently have only two digits, why don't you experiment by changing things like the value of **DeltaVal** (eg, making it +2 or -3)? Or how about changing **MIN_SCORE** to be -9 and **MAX_SCORE** to be 42, for example?

Next time

In our next column, we will add the remaining digits to our display and incorporate a full complement of these modules into our games console. Furthermore, despite previous broken promises, we really will add the eight push-button switches to our console's front panel.

As always, if you have any thoughts you'd care to share on anything you've read here, please feel free to email me at max@clivemaxfield.com

Devices and components used in this issue

Arduino Uno R3 microcontroller module	https://pemag.au/link/ac2g
Solderless breadboard	https://amzn.to/3O2L3e8
4-inch (10cm) jumper wires (optional)	https://pemag.au/link/ac2l
8-inch (20cm) jumper wires (male-to-male)	https://amzn.to/3O4hnxk
LEDs (assorted colours)	https://amzn.to/3E7VAQE
Resistors (assorted values)	https://amzn.to/3O4RvBt
Ceramic capacitors (assorted values)	https://pemag.au/link/ac2i
16V 100µF electrolytic capacitors	https://pemag.au/link/ac2j
8-Segment DIP red LED bar displays	https://pemag.au/link/ac2c
74HC595 8-bit shift registers	https://pemag.au/link/ac1n
Dual 7-segment display modules	https://pemag.au/link/ac8i
0.1" pitch header pins	https://pemag.au/link/ac8j

Putting the boot in

I'm constantly thinking of things I want to tell you. Are these things necessary for your continued existence? Probably not, but—in my experience—the more we know about how things work, the happier we are at the end of the day.

In this vein, a 'bootloader' is a small program stored in a microcontroller's flash memory. The name comes from 'bootstrap loader' (as in "pulling yourself up by your bootstraps"), because it's the very first piece of code that runs when the chip powers up or resets. Its job is to 'boot' the system by loading or launching the main user program.

The Arduino Uno ships with a pre-installed bootloader that occupies 512 bytes of the Arduino's 32,768-byte (32kiB) flash memory. This bootloader enables the chip to receive new programs over the USB-serial link without requiring an external hardware programmer.

If the Arduino detects that the Integrated Development Environment (IDE) on your main computer is attempting to upload a program, it writes that program into flash memory; otherwise, it quickly jumps to whatever existing user program is already stored in flash.

The Fox Report

Barry Fox's technology column

A blurry, indistinct photo isn't 'proof of delivery'

If you buy goods online – and who now doesn't, even if they would prefer to shop in person – this is for you. Keep it in mind when you next consider ordering from any seller using Royal Mail Tracked as its carrier service.

Royal Mail's letter service is legendarily unreliable. To my acute embarrassment, a birthday card I sent weeks ago still has not arrived. But Royal Mail Tracked delivery is supposed to be different, with real-time, end-to-end tracking of items along with photo proof of delivery. It's the service online sellers pay extra to use (and charge us extra for).

Recently, for the second time, electronic components I ordered from an eBay seller were sent by Royal Mail Tracked and failed to arrive. In each case, the (different) sellers provided a meaningless photo as a worthless 'proof of delivery'. In each case, the eBay seller failed to challenge the useless proof from Royal Mail, telling me to contact them.

I quote: "We already clearly tell to you that Your item was delivered..from our side we cant do nothing please contact with the courier service".

This is wrong; the seller is Royal Mail's customer, having paid them to carry the package. Royal Mail's official advice note headed "Marked as delivered but not received" clearly tells recipients who have failed to receive goods: "If you're the recipient. Please contact the sender. They will contact us to enquire on your behalf". See **https://help.royalmail.com/personal/s/article/Marked-as-delivered-but-not-received**

I know from experience that it is relatively easy to sideline an unhelpful seller and make a chargeback request with the credit card company used to pay for goods or services. The card company considers the facts and claws back money from the seller if things look wrong.

But as an exercise for these pages, I hacked through eBay's jungle of looped FAQ Help pages, chatbot advice and no-reply messaging to pursue a complaint and help others with the hassle.

I've itemised the bare bones here, but note that eBay's menu options may change depending on what steps you have already taken. Also, only correspond with the seller via eBay. Do not contact the seller directly, even if you have an email address.

Open Your eBay Account, go to My eBay, Purchase History, find the item in question, click "More Actions" and use the "Contact Seller" option to write politely to the seller, confirming non-receipt.

If the seller isn't helpful, go again to Purchase History/More Actions, find the same item under Purchases. Then select "See Request Details" and "Ask eBay to Step In and Help". Choose the appropriate reason and Confirm.

You will (quite reasonably) need to assure that you have checked that the delivery address on the order was correct and have checked for misdelivery with neighbours and in possible 'safe places'.

Having done this, I complained that the fuzzy tracking photo provided showed no signature, no person and no house, and was thus obviously not proof of delivery. I also complained that the seller had wrongly told me to contact Royal Mail.

Nevertheless, eBay rapidly rejected the complaint in a brusque "Case Is Now Closed" email. This offered no reason or reasoning and had a distinct whiff of Artificial Unintelligence automated responding about it:

"This case is now closed... the seller provided tracking information that shows the item was delivered to your address. After reviewing all the details of this case, we determined that you won't receive a refund."

There was no obvious option to appeal eBay's unreasoned ruling. Still, the option exists. The trick is to access the Case Outcome details, which contain a small-print link named "If you disagree with the outcome, you can send an appeal". But finding the Case Outcome is not easy, especially as the menu options change with case progress.

One way is to search Customer Help for "Appeal the outcome of a case as a buyer" and click on the link offered; or you can go to Purchases/See Case Details and look for the option to "Send us an appeal". Condense your text to the short length limit and Submit.

eBay warns that it will only consider appeals against its decisions if there is new evidence. But just for the heck of it, I appealed the refusal to refund, essentially re-stating the facts and adding that "eBay has failed to give reasons for relying on self-evidently worthless 'proof of delivery'... This undermines eBay's guarantee promises... I no longer dare order anything from eBay that involves Royal Mail Tracked delivery".

Expecting another rejection, I started preparing a chargeback request to my credit card company. This is the process where the card company that made the original payment considers the facts of a dispute and claws back money from the seller, or middleman (eBay in this case), if they believe it is valid. The burden of proof then shifts to the seller, or eBay.

The usual routine is for the seller to make a token defence by parroting what they already said, but give up when the cardholder rebuts.

As if psychic, eBay Customer Service then came back with a rethink:

"After reviewing this case, we decided to refund you for this purchase. We did this because we determined that the item was stolen during delivery."

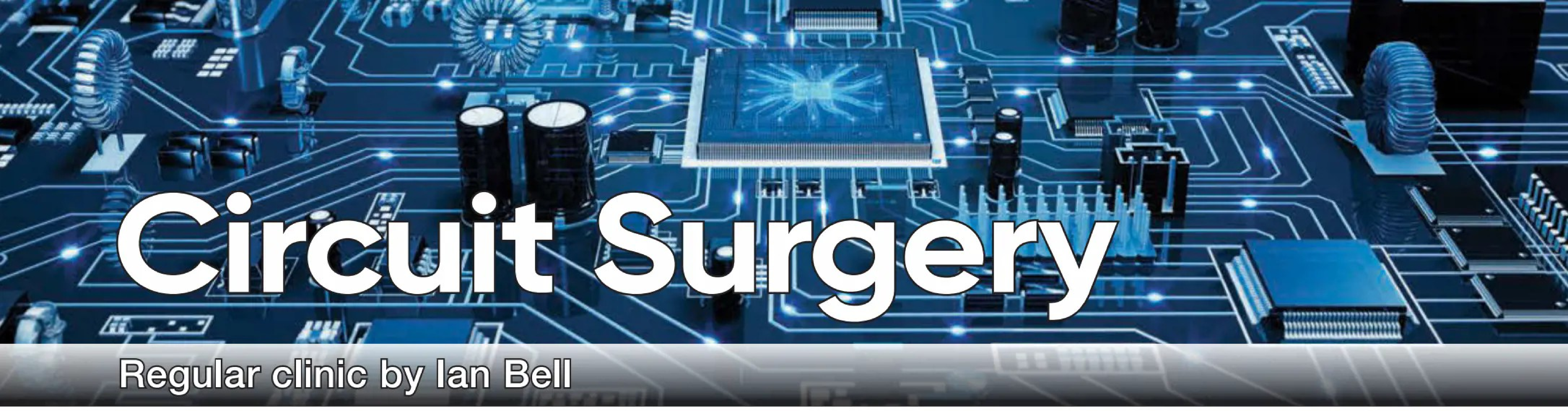
In a separate email, an eBay employee (or AI-created character?) added: "I understand that the tracking shows item has been delivered but you have not received the item. … we've granted your appeal and as a courtesy, I'm happy to issue this refund for you. … Please note, this is a one-time exception to our policy and we will be unable to issue this courtesy again if future situations like this arise. … Have a wonderful day."

So, eBay was not admitting any mistake or failure in not challenging obviously worthless "proof of delivery". Most significantly, eBay then offered "Some useful tips that can help to ensure that your future deliveries arrive safely." Apart from the obvious "ensure that the correct address is on the order details when purchasing", these tips are:

- Select Click & Collect for hassle-free delivery to a collection point near you or
- Choose a secure delivery location (like a locked porch box or an indoor delivery if available) or
- Redirect the package to a nearby pickup location, or locker service, or by using eBay's Click & Collect Service.

This appears to be clear advice **not** to buy from any seller who only dispatches goods by Royal Mail Tracked. It's exactly the conclusion I had already reached. I shall in future buy from Amazon, which uses its own courier service; or routinely avoid any transaction that involves carriage by Royal Mail Tracked.

With customer advice like this from a sales organisation as big as eBay, how long can Royal Mail stay in the mail business?



Circuit Surgery

Regular clinic by Ian Bell

Measuring the frequency response of a circuit using a PC sound card, part 3: using op amps for signal conditioning

Last month, we continued looking at how to measure frequency responses. That followed from the article before, where we concentrated on the principles and theory of frequency responses of linear circuits and how to measure them.

In the follow-up article last month, we described how to use the sound card (audio interface) of a PC or laptop together with free software called REW (Room EQ Wizard) for practical frequency response measurement. REW is aimed at acoustic measurements (hence the name) but can be used to analyse purely electronic circuits as well.

We already satisfied our objective of measuring the frequency response of an example digital filter from the recently completed DSP series. However, we did not have space to go into details of the circuitry used to interface the DSP filter to the PC's audio I/O.

The processes that manipulate signals so that otherwise incompatible analog circuits/stages can pass signals between them is referred to as 'signal conditioning'.

Signal conditioning commonly requires amplifiers and attenuators, sometimes with variable gains, and these are often implemented using op amps. So this month, we will look at op amp amplifiers, particularly for AC signals, covering a variety of general circuit design and component selection issues, as well as the specific circuits for the frequency response measurements described last month.

Before getting into the op amp basics, we will recap the context of the circuitry used to assist the sound-card-based measurement.

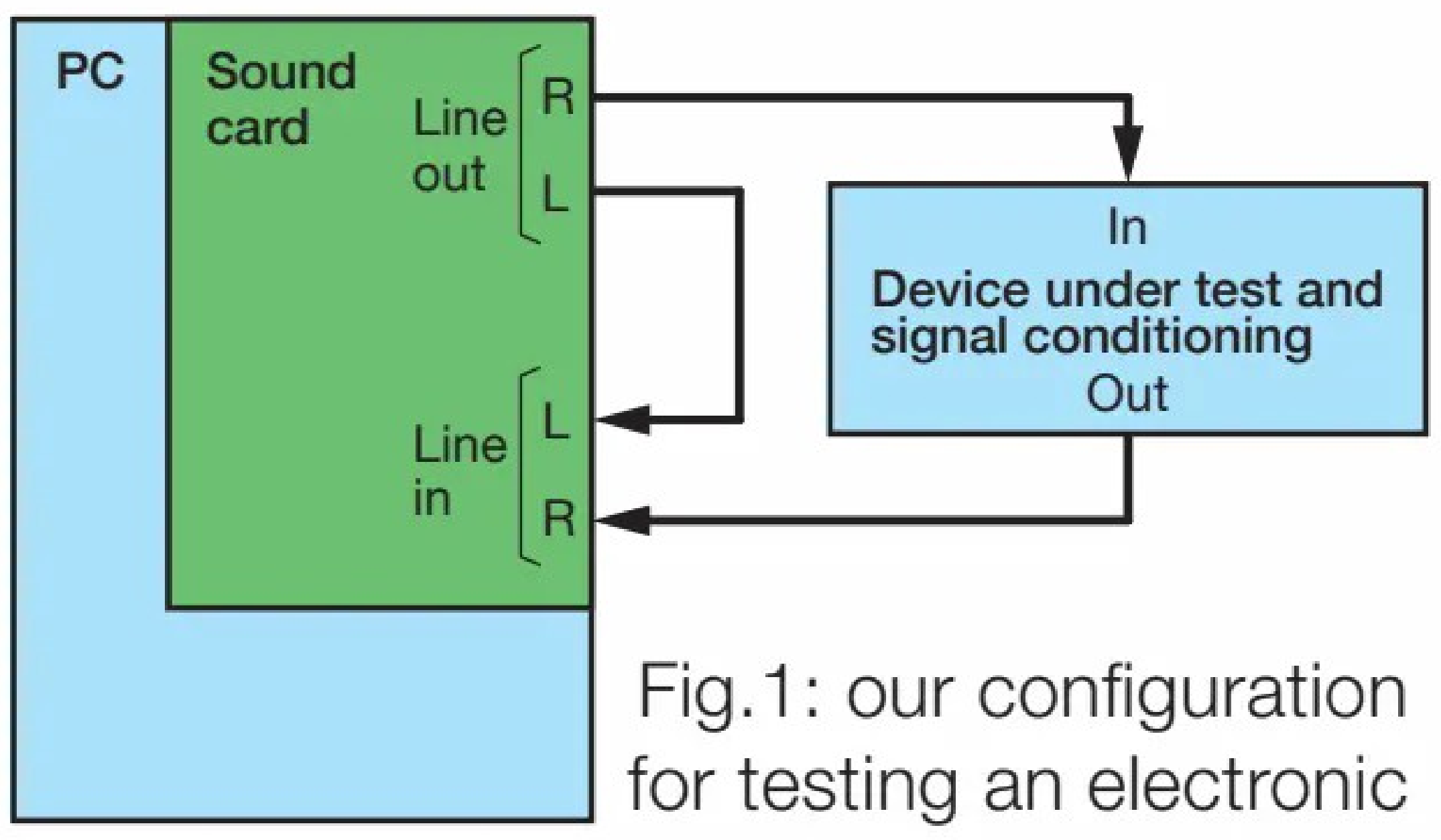


Fig.1: our configuration for testing an electronic device using a computer sound card.

Our measurement system

The basic measurement setup is shown in Fig.1. One channel of the sound card's stereo line (or headphone) outputs is used to produce the input signal to the device under test. The output from the device under test is fed into one channel of the sound card's line input.

REW (or similar software) generates the signal used for testing and processes the resulting response to provide real-time displays of signal levels, waveforms (oscilloscope function) or the signal spectrum. The software can also plot graphs of frequency response and distortion vs frequency and other parameters by running a measurement process.

The other stereo channel can be used as a timing reference via a direct output-to-input (loop-back) connection.

Computer audio interfaces have a limited range of output voltage, typically close to either the commercial or professional standard audio line levels, depending on the type of sound card; in either case, it is generally in the order of 1V RMS.

The audio interface can also only process AC signals centred on 0V, so unlike lab test equipment, DC voltages cannot be output or measured. As a result, the signal amplitudes and DC offsets of the sound card may not match those required by the device under test.

We may need amplification/attenuation and DC level shifting at both the input and output of the device under test. Furthermore, even if the voltage levels are compatible, the sound card output may not be able to drive the circuit under test (and vice versa) due to loading effects – the output could be overloaded, or the load may cause instability in the driving circuit.

Last month, we discussed in general terms the signal conditioning circuit used for the frequency response measurement of the example digital filter. This structure of the circuit is shown in the block diagram, Fig.2. The circuit could also be used for a range of different devices under test, not just the DSP system, so some circuit parameters, such as gains, may need to be different.

We previously discussed the specific signal levels and gains required for interfacing the digital filter to the sound card I used. To recap briefly, the first buffer AC-couples the output from the sound card and has a gain of 2.6 times to obtain 3.2V peak-to-peak (for a sinewave) at the filter's input from the 1.22V peak-to-peak maximum output from the sound card.

The second buffer has a gain of about 0.19 times to obtain 0.61V peak-to-peak from the 3.2V peak-to-peak filter output, which is amplified to 1.22V peak-to-peak by the 2× gain of the audio driver to give an overall unity gain between the sound card input and output.

The DSP system under test also has some signal conditioning at its input to shift the signal to a DC level of +1.65V (half the microcontroller's ADC reference voltage) and provide input over-voltage protection (see the August and September 2025 issues). There is also AC-coupling and a reconstruction filter on the DAC output.

Op amp based amplifiers

The signal conditioning circuitry was built using op amps. This month, we will discuss aspects of op amp circuit design relevant to the two buffers shown in Fig.2.

Figs.3-5 show three widely used op amp circuits. Fig.3 is an inverting amplifier, with a gain of $-R_F \div R_I$. The inverting amplifier has a phase shift of 180°, which flips the waveform upside down, hence the name.

The subscripts are F and I for feedback and input resistor, making it easier to

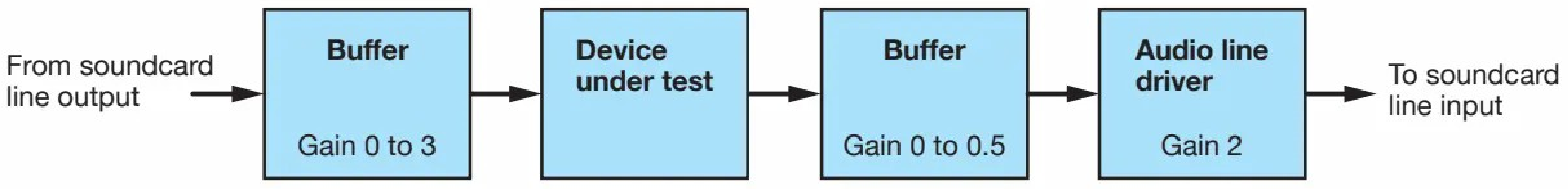


Fig.2: the signal conditioning circuitry for testing the DSP filter.

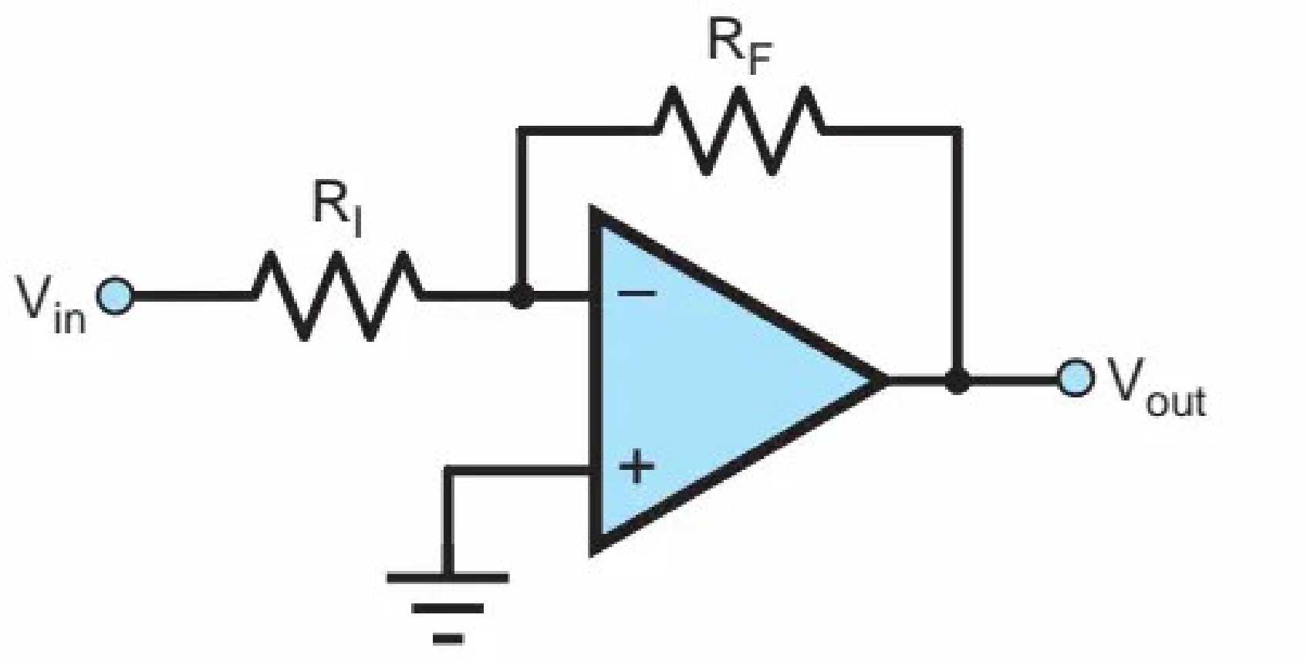


Fig.3: an inverting op amp amplifier.

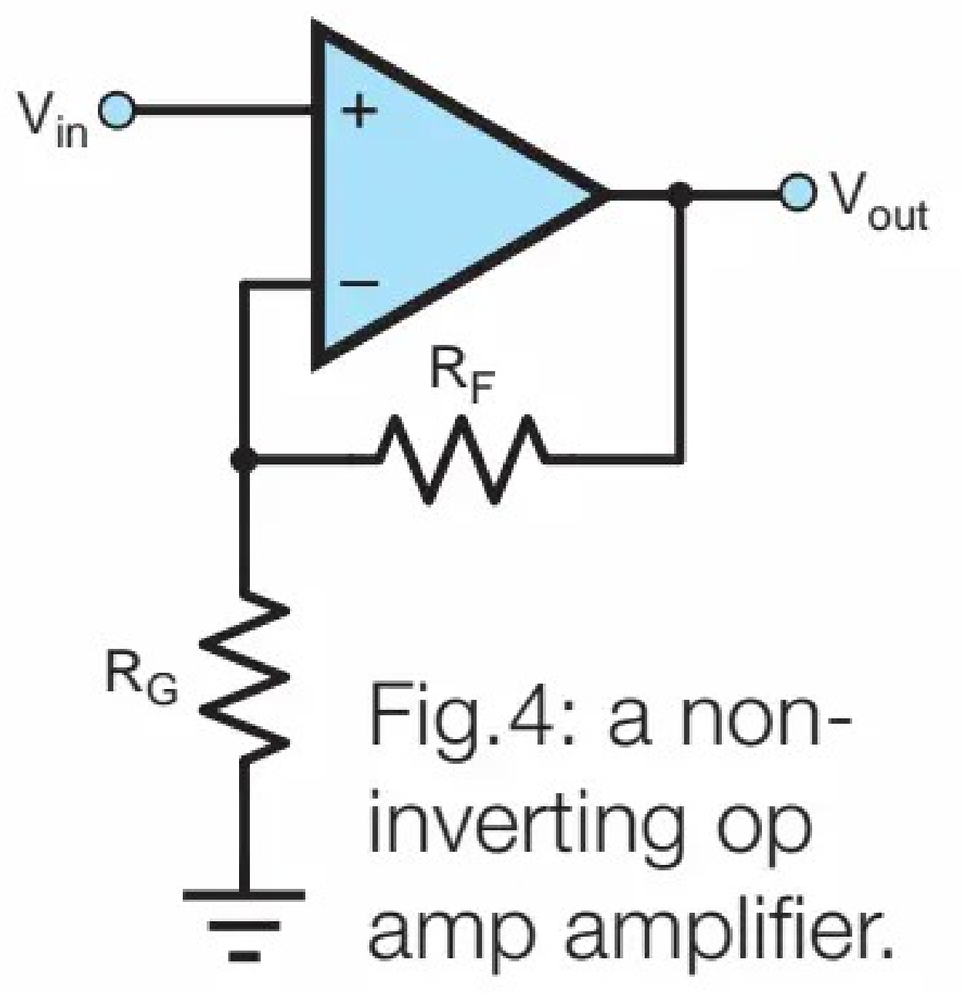


Fig.4: a non-inverting op amp amplifier.

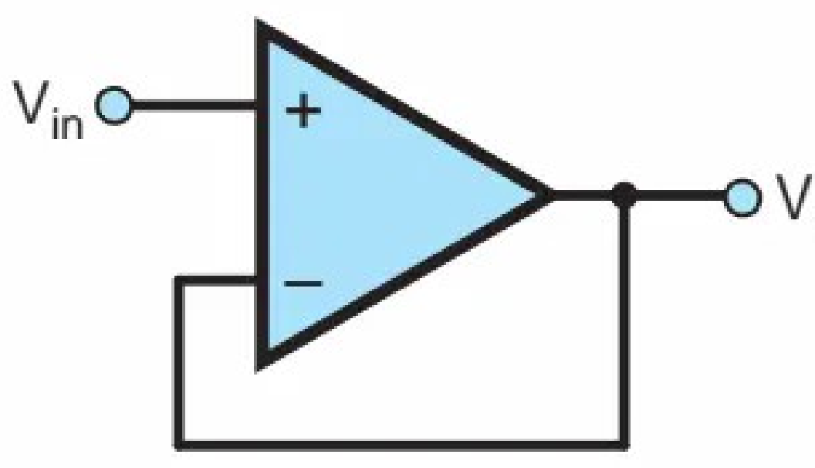


Fig.5: an op amp based buffer (gain=1).

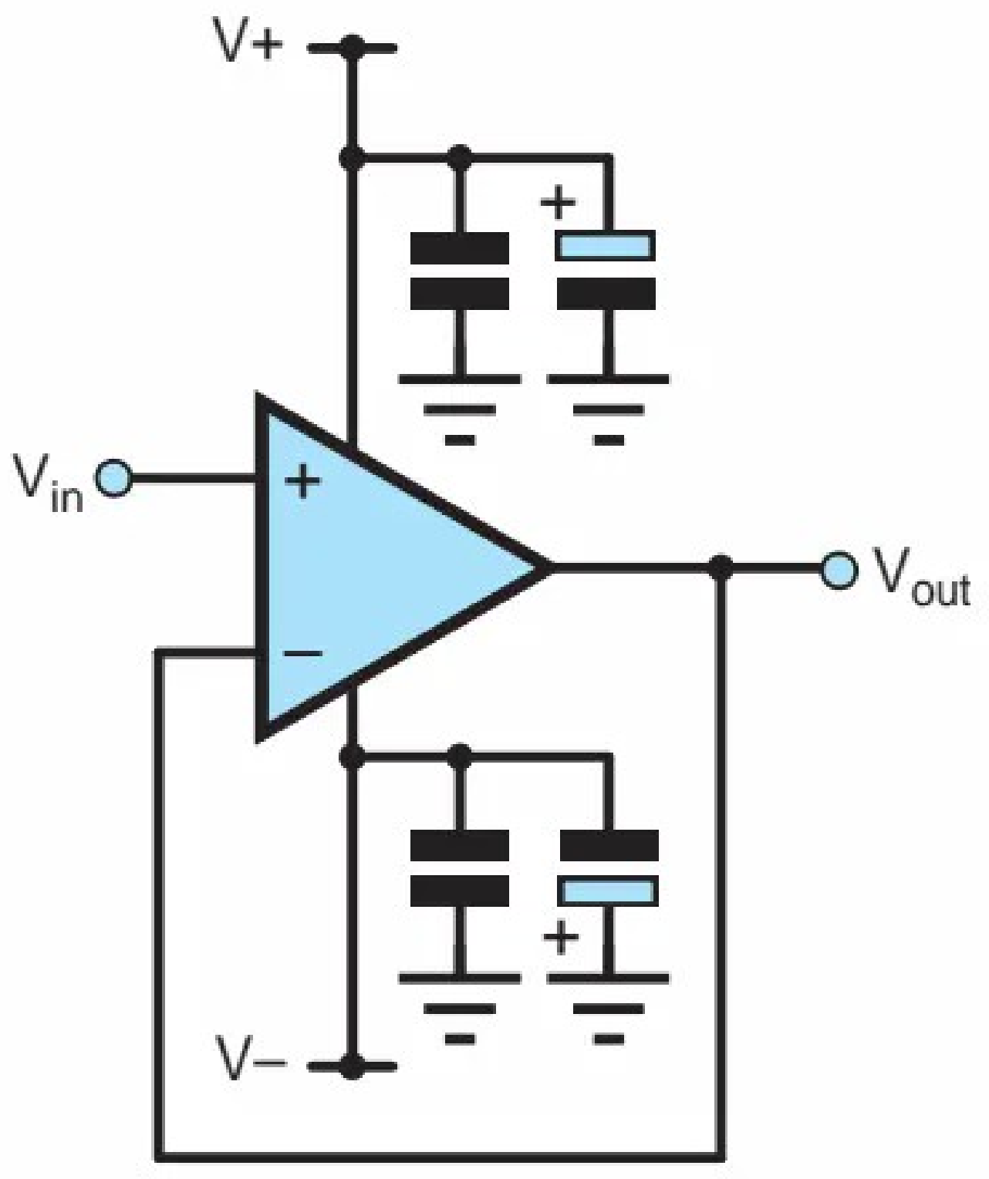


Fig.6: an op amp powered by split supplies, with suitable supply bypass capacitors.

remember the gain formula than if you designate them R_1 and R_2, where it is not necessarily obvious which is which.

Fig.4 shows a non-inverting amplifier, which has a gain of 1 + $R_F \div R_G$ (G is for grounded resistor) and zero phase shift. That equation is equivalent to ($R_F + R_G) \div R_G$.

Fig.5 shows a unity gain buffer, which is a version of the non-inverting amplifier with $R_F = 0$ and hence a gain of 1.

Power supplies and bypassing

The circuits in Figs.3-5 do not show the op amp's power supply connections. This is common practice when discussing circuit configurations, as it keeps the schematics simple, but of course the op amps won't work without suitable supply rails.

Op amp circuits are often run from split supplies, ie, positive and negative voltages of the same value. For example, ±5V, ±12V or ±15V. This generally leads to simpler circuits, particularly when dealing with AC signals centred on 0V, because the circuit can handle positive and negative signal voltages without any DC level shifting.

The signal conditioning circuit for the example digital filter uses split supplies.

Some op amps have inputs and outputs that can operate almost at their supply rails and such an op amp with a ±5V supply would be able to handle signals of almost 10V peak-to-peak (3.5V RMS for a sinewave) without problems.

However, keep in mind that many common op amps can't do this, and have limitations on the input or output signal swing (often both). An op amp like the NE5532 (which is old, but inexpensive and performs very well) can only handle a signal of around ±3V with ±5V supply rails (2.12V RMS).

If this is insufficient for your application, you need higher supply voltages (up to the maximum the op amp can handle) or a 'rail to rail' op amp. Some op amps have a rail-to-rail output but not input (eg, the TL971/2/4 series), which is not helpful if you are using them as a buffer but works well in circuits with a gain above unity.

Others, like the TSV991/2/4 series, have inputs and outputs that can swing to both rails (rail-to-rail input/output or RRIO), so they can buffer signals that swing between either supply rail. The trade-off is that they usually cost more, can be noisier and usually don't have as low signal distortion figures.

Power supplies are not perfect, so there will be unwanted voltage variations over a wide range of frequencies, which may include noise from power supply circuitry and perturbations caused by varying voltage drops across the supply resistance and inductance as the circuit's supply current changes. This is called power supply noise.

Circuits such as amplifiers will ideally ignore power supply noise, so it does not affect their outputs, but of course they are not perfect. Op amp data sheets may state a power supply rejection ratio (PSRR) which indicates how good they are at this. They usually have excellent rejection at lower frequencies (most of the audio range) but at higher signal frequencies, some noise and ripple may creep through.

The impact of power supply noise can be mitigated by placing capacitors across the power supply – this is called power supply bypassing.

The bypass capacitors can be thought of as shorting out high-frequency noise – the effective resistance of an ideal capacitor decreases with increasing frequency, so bypass capacitors short the supplies together at high frequencies and shunt the noise past the circuit. No direct current flows into the capacitors once they have charged to the supply voltage at power-up.

Bypass capacitors are also required for stability – op amps without them can act as oscillators, which is usually not what you want! So don't forget to include them for every op amp package.

Typically, the connections from the power supply to the op amp may be quite long (eg, wires inside an enclosure, or from a bench power supply to a prototype, followed by traces on a PCB or breadboard connectors). The wiring is not ideal, not a perfect conductor, so it has unwanted (parasitic) resistance and inductance.

When the supply current changes abruptly, the parasitic inductance of the wiring will oppose the change, resulting in a voltage drop proportional to the rate of change. Effectively, the wiring parasitics impede the flow of charge from the supply to the op amp. Capacitors across the supply closer to the op amp provide a reservoir of charge, which can respond more quickly than the power supply via the longer wiring.

Like the supplies, capacitors are not perfect, so the use of just one type and value capacitor is not fully effective at supply bypassing. A relatively large capacitor (eg, a 10µF to 100µF electrolytic) can provide a good charge reservoir and deal with low frequency noise, but its own parasitics (internal inductance and resistance) make it less effective against high frequency noise.

To cover this, smaller capacitors (eg, 10nF to 100nF) are also commonly used. The two types are connected in parallel across the supplies. For split supplies, the capacitors can be connected from each supply to ground, as shown in Fig.6. However, this is not always necessary, so if testing shows good performance is achieved with capacitor(s) between the supply rails, the parts count can be reduced.

Commonly for split supplies, a pair of 100nF capacitors would be placed as close as possible to each op amp package, and a single pair of larger capacitors would cover a board or group of chips up to a distance of a few centimetres. Device data sheets may provide specific recommendations on supply bypassing.

In a little more detail, the benefits of capacitors between V+ and V- are more charge storage (energy is CV^2 and here V is doubled, so energy storage is quadrupled), lower parts count, and better bypassing of the actual supply paths within the op amp (with a split supply, the op amp package has no direct connection to ground anyway).

The main advantage of the pair of capacitors, V+ to GND and V- to GND, is that almost all op amps use either V+ or V- as a common terminal for their internal amplifier (usually V-). Therefore, they are more sensitive to noise on that terminal. By bypassing it individually to GND, you effectively make that rail an AC ground, improving the op amp's PSRR.

Which approach is best really depends on the op amp and how it is being used, so if you need the best performance, you may need to experiment.

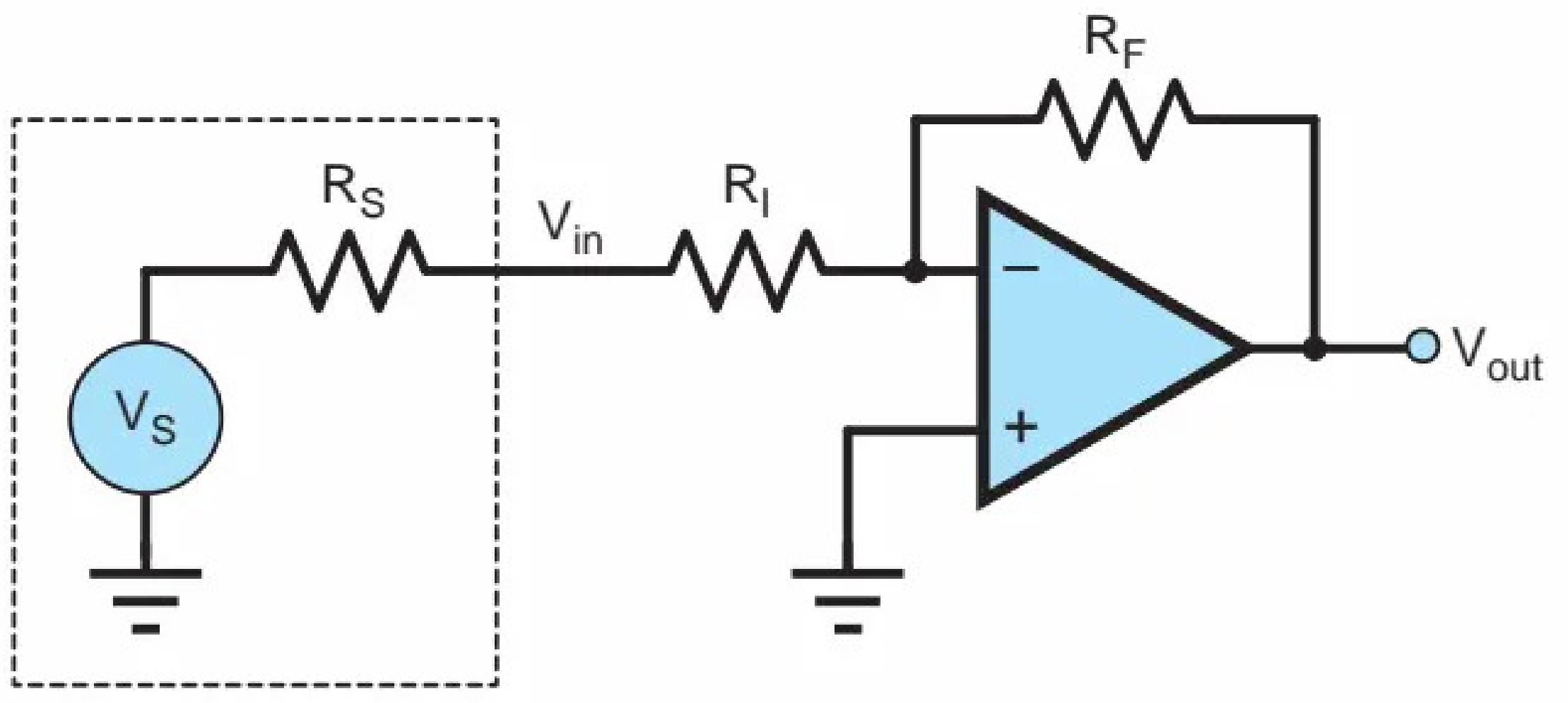


Fig.7: an inverting amplifier with its signal source.

Inverting or non-inverting?

If we need amplification (a gain greater than 1) we can use either the non-inverting or inverting circuit – so how do we choose between them? The circuits have the same number of components, and we can set any reasonable gain greater than one using suitable resistors, so there is no significant difference there.

On the other hand, if we specifically need to invert, or avoid inverting, the signal, the choice may be straightforward. In some cases, the non-inverting amplifier will be preferred simply because it does not invert the signal, for example, where phase relationships between multiple channels or signal paths are important.

The non-inverting circuit has a minimum gain of one (unity), so it cannot be used to directly implement an attenuator, or variable gain that varies from cut to boost. However, it is straightforward to implement attenuation using a potential divider. The inverting configuration can be used as an amplifier or attenuator.

There are significant differences between the input impedances of the two types of amplifier, and somewhat related to this, differences in the consequences of AC coupling their inputs. We will look at this in detail shortly. There are also subtle differences in noise and distortion performance between the two configurations, which we will discuss later.

When we go beyond amplifiers to other uses such as summing circuits and filters, different arguments may be relevant to whether the inverting or non-inverting configuration should be employed. In general, it depends on the application, so our discussion here will mainly highlight issues related to AC amplifiers.

Virtual earth/ground

The input impedance of the inverting amplifier is equal to R_I. This is a consequence of the way the feedback operates in the circuit. The negative feedback used in op amp circuits ensures that the voltage difference between the op amp's inputs is close to zero as long as the op amp is not pushed beyond its limits.

The amplifier circuits are effectively control systems that 'try' to maintain zero voltage across the inputs. Zero volts between two points is also what you get in a short circuit, so the op amp's inputs behave almost like they are shorted together – referred to as a virtual short circuit.

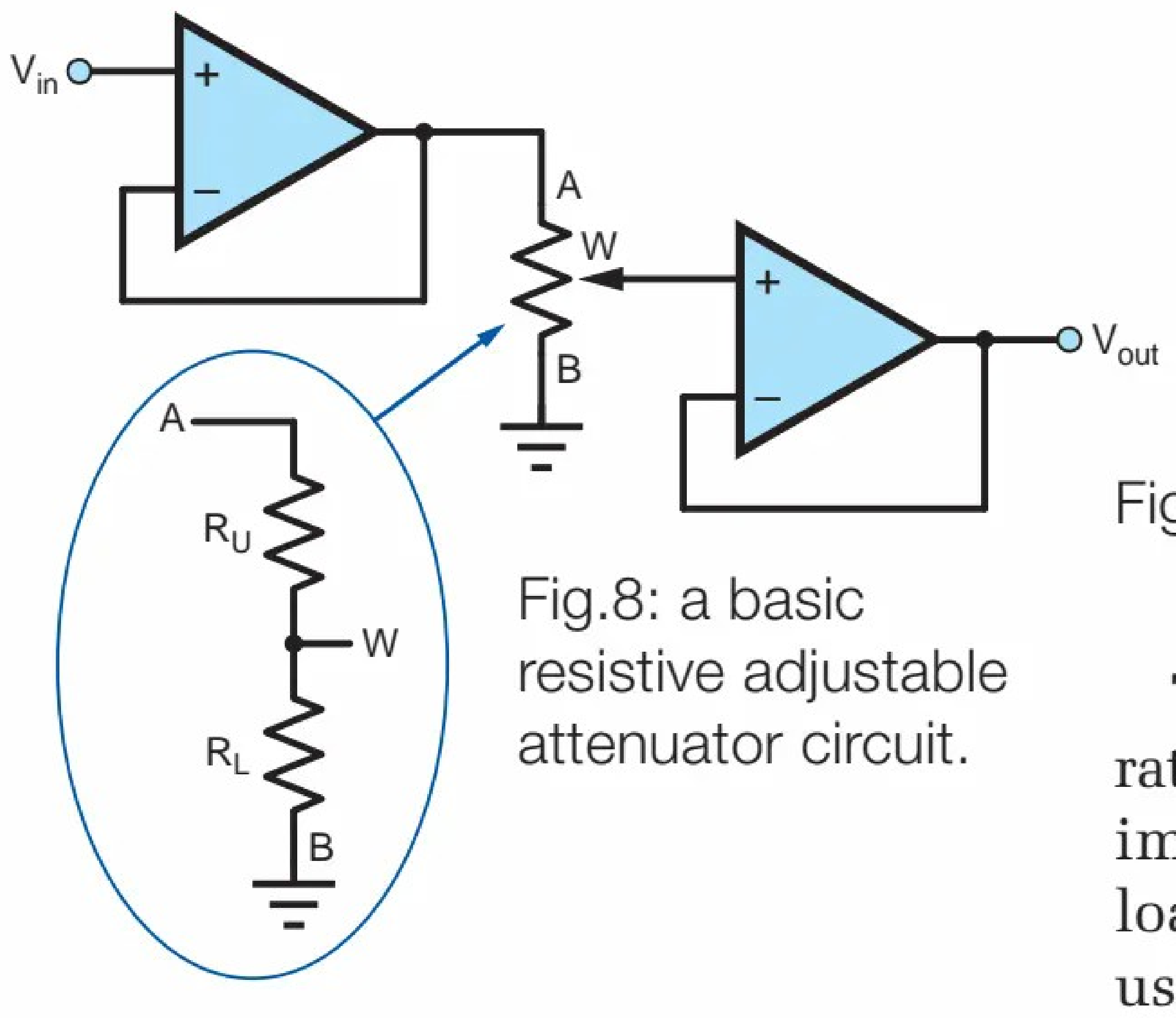


Fig.8: a basic resistive adjustable attenuator circuit.

We say 'almost' because the op amp is not perfect; ideally, it has infinite gain, but in practice the gain is finite, so the 'short circuit' is not perfect.

The inverting amplifier has its non-inverting input connected to ground, so the inverting input behaves as if it is shorted to ground – referred to as a virtual earth or virtual ground. R_I is connected from the input to the virtual earth, so the input impedance is equal to R_I.

This typically means that a very high input impedance cannot be obtained with the inverting configuration, because using very large resistors to set the gain has some undesirable consequences.

These potential problems include:

- higher thermal (Johnson) noise from larger resistor values
- higher sensitivity to external interference; higher impedance nodes in the circuit are more susceptible to picking up unwanted signals
- more likelihood of instability due to the effects of stray capacitance (for a given capacitance, unwanted phase shift in the feedback is larger with larger resistors)
- larger DC errors due to bias currents flowing in the resistors (less important with AC-only signals, but may lead to asymmetric clipping)

Source loading

Fig.7 shows the inverting amplifier connected to a source voltage (v_S) with output resistance R_S. The source resistance and R_I form a potential divider. The voltage at the amplifier's input (v_{IN}) is ideally equal to v_S, but is reduced to $v_{IN} = (R_I \times v_S) \div (R_I + R_S)$ by the loading imposed by R_I (this is the well-known potential divider equation).

Alternatively, we can say the effective gain (with respect to v_S) is reduced by the same factor.

The non-inverting configuration has a very high input impedance as the input is connected directly to the op amp's input. Op amps typically have an input impedance of megohms, gigaohms or even teraohms.

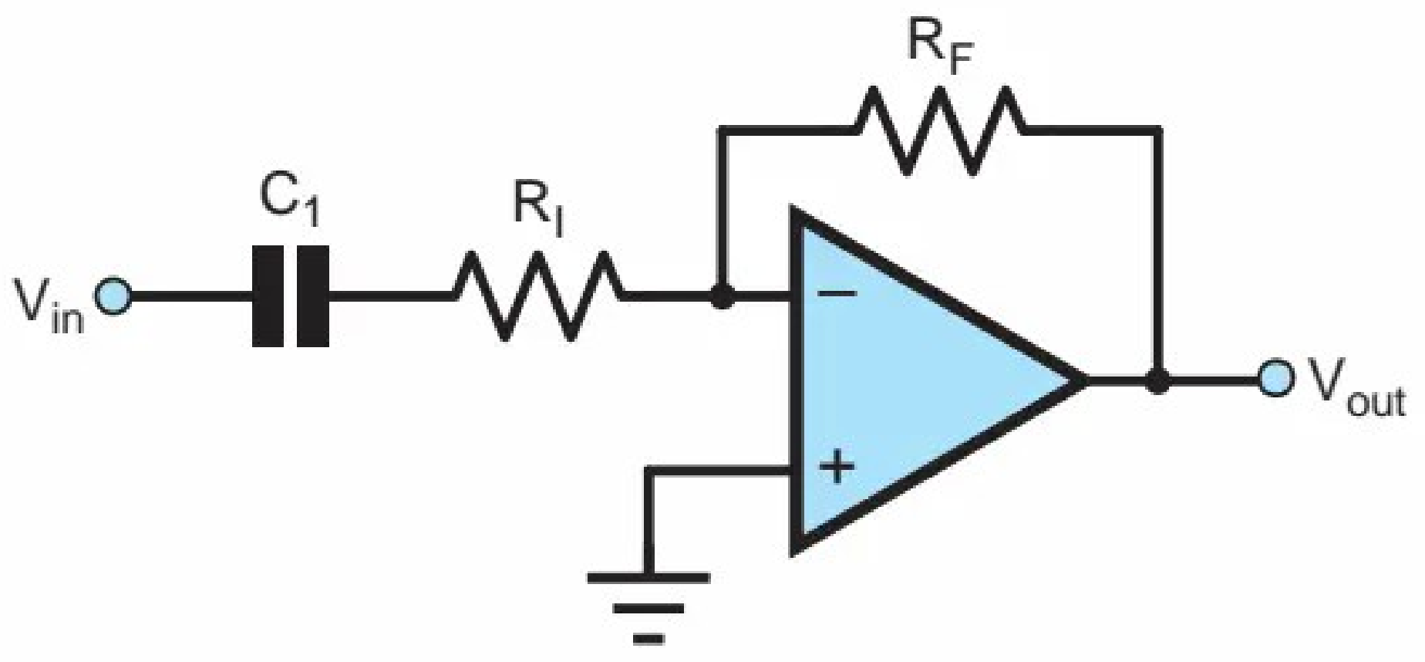


Fig.9: an AC-coupled inverting amplifier.

This makes the non-inverting configuration very useful for connecting to high impedance signal sources, where the loading effect would be excessive when using an inverting circuit with typical resistor values, or where the source impedance may be variable or unknown, which would otherwise result in uncertainty or variability in the effective gain.

The unity-gain buffer is used where gain is not required but signal loading is a potential problem at a system/stage input, or for driving loads that have too low an impedance for a system/stage output.

Potential divider attenuator

As mentioned above, the non-inverting amplifier cannot provide a gain below unity (ie, no attenuation). The circuit shown in Fig.8 has a potentiometer between two buffers, which provides a variable attenuation from 0× to 1×. If a fixed attenuation is required, a fixed resistor potential divider can be used, as shown.

Resistors can be used in combination with a potentiometer to restrict the attenuation range and give finer control.

The op amp circuits do not have to be unity gain buffers – the potentiometer could be driven by the output of any op amp stage, for example, a preceding amplifier or filter. An input buffer is required if the potentiometer is to be connected to a relatively low source resistance, where it will cause loading.

The op amp at the output of the potentiometer can be any circuit with a high input impedance (typically one based on the non-inverting configuration).

In audio design, it is common practice to AC-couple volume control potentiometers because DC on the potentiometer could (likely will) cause unpleasant crackling noises due to track irregularities as the wiper is moved, resulting in steps in the output voltage.

For the frequency response measurement circuit, this is not a concern because there is not the same need to provide a 'good listening experience'; we are using trimmers to adjust gain to what is likely to remain a fixed value for a given measurement setup. Furthermore, the buffer circuits will be processing purely AC signals due to AC coupling earlier in the signal path, which brings us to the AC-coupling of op amp amplifiers.

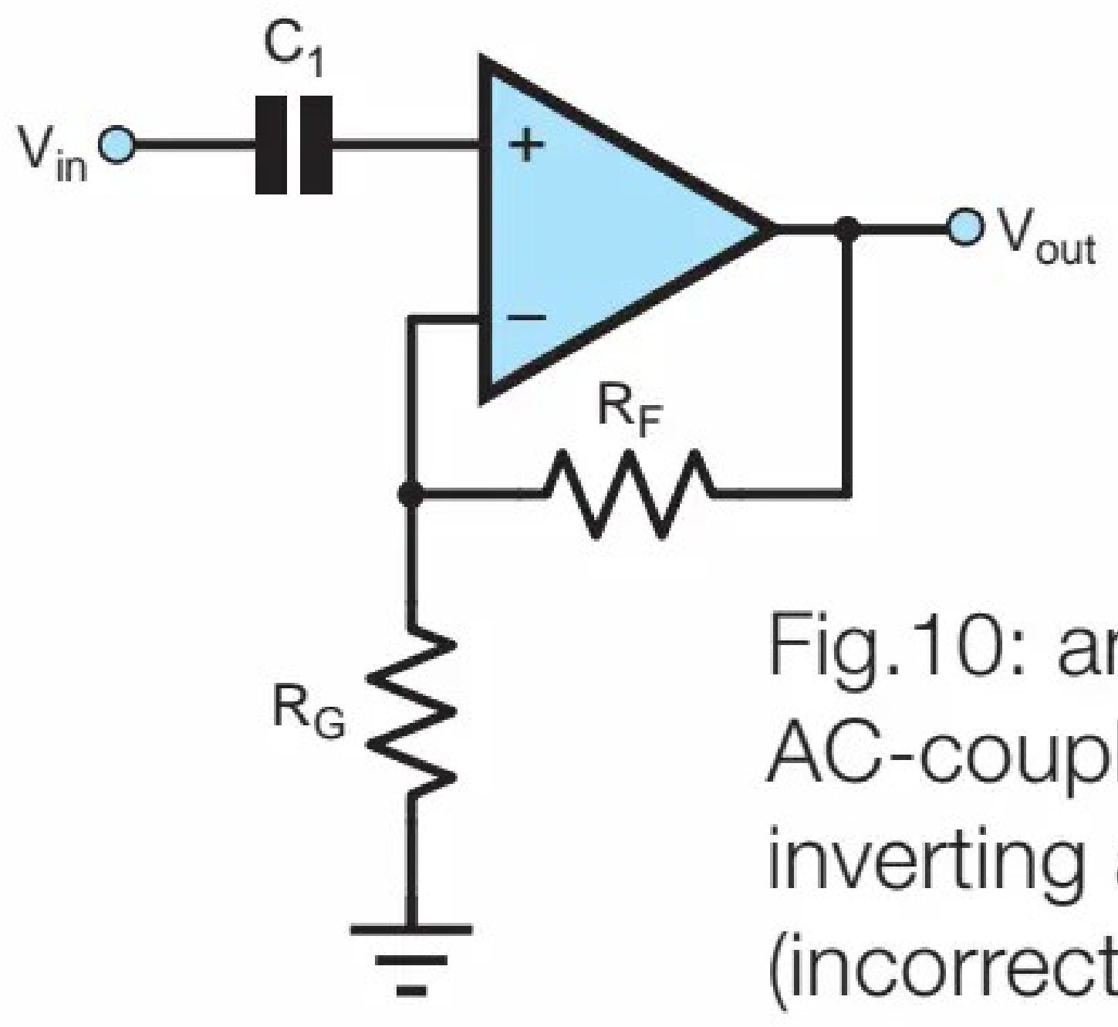


Fig.10: an AC-coupled non-inverting amplifier (incorrect version).

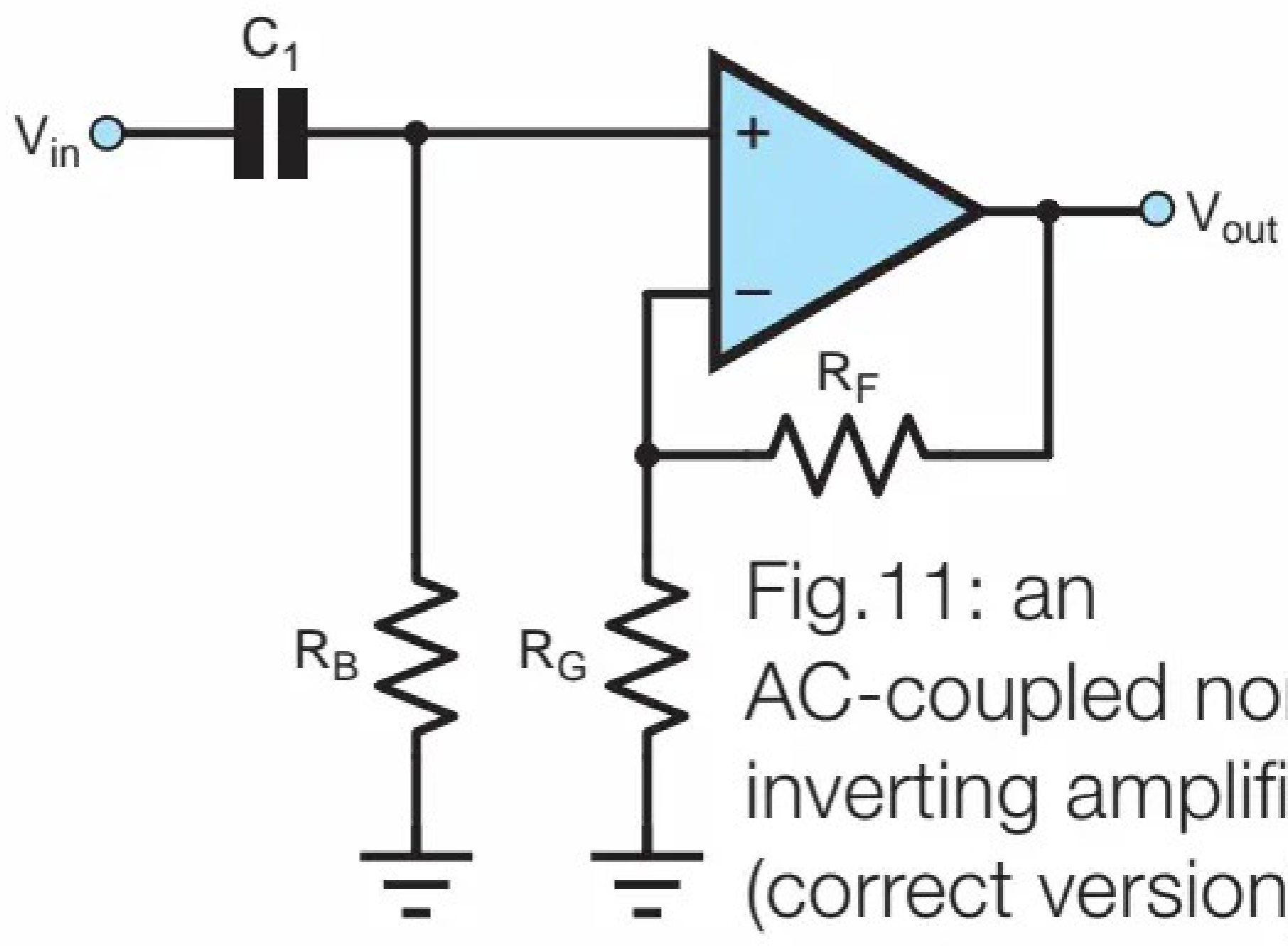


Fig.11: an AC-coupled non-inverting amplifier (correct version).

AC coupling

For the circuits in Figs.3-5, we can assume (in general) that the signals (v_{IN} and v_{OUT}) are a combination of DC and AC. The gain equations apply to DC and AC as long as the op amp's operating range is not exceeded. If we want to process AC signals only, we typically AC-couple the input using a capacitor. This is shown in Fig.9 for the inverting amplifier.

The coupling capacitor attenuates the signal at low frequencies as well as blocking DC. The reactance (effective resistance) of the capacitor (X_{C1}) increases the effective value of the input resistor, reducing the gain magnitude to $-R_F \div \sqrt{R_I^2 + X_{C1}^2}$.

The effective value of the input resistance is the magnitude of the complex impedance of the series combination of R_I and X_{C1}. Due to the phase-shifting properties of the capacitor, we cannot simply add R_I and X_{C1} – the effective resistance is obtained using the root-mean-square value, as shown above.

We can apply the commonly used -3dB value to define the low-frequency cutoff (f_c) due to inserting the capacitor. -3dB is used because it is the point at which output power falls by half. As voltage is proportional to the square root of power, half-power corresponds to the voltage decreasing by a factor of $\sqrt{2}$ (0.7071).

The gain of the inverting amplifier reduced by $\sqrt{2}$ can be written $-R_F \div \sqrt{2}R_I$. Combining this with the equation above, we get $\sqrt{R_I^2 + X_{C1}^2} = \sqrt{2}R_I$, as the R_F terms cancel. Squaring both sides gives $R_I^2 + X_{C1}^2 = 2R_I^2$, so $X_{C1}^2 = R_I^2$ (ie, $X_{C1} = R_I$) at the -3dB point.

The reactance of a capacitor of value C at frequency f is $X_C = 1 \div (2\pi fC)$, so the cut-off frequency at which $X_{C1} = R_I$ is obtained by rearranging $R_I = 1 \div 2\pi f_c C_1$ to give $f_c = 1 \div 2\pi R_I C_1$.

This is the same as the cut-off frequency of a high-pass RC filter. We could have just assumed this would be the case, but as the circuit topology is not exactly the same, it is worth analysing.

Rearranging the cut-off equation to find $C_1 = 1 \div 2\pi f_c R_I$ allows us to choose a suitable AC-coupling capacitor once we have selected the gain-setting resistors.

Non-inverting amplifiers

Given that AC-coupling generally means inserting a capacitor between the signal source and circuit input, it may seem that an AC-coupled version of the inverting amplifier should be as shown in Fig.10. Unfortunately, things are not that simple, and the circuit in Fig.10 will fail to operate correctly. This is because op amps require DC bias currents at their inputs.

The rule for op amps is that you must provide a DC path to both inputs, but capacitors block DC, so the circuit in Fig.10 fails to meet this requirement.

For the circuit in Fig.10, the op amp will still take its bias current (I_B), which for simplicity we can assume to be a constant current. This could be in either direction (in or out of the op amp) depending on the device's internal circuitry and operating conditions.

When a constant current flows into or out of a capacitor, the capacitor will charge or discharge. Thus, in the circuit in Fig.10, the coupling capacitor will charge from the bias current.

When a capacitor charges (from zero charge), the DC voltage across it increases, with the polarity of the voltage determined by the current's direction. Assuming $v_{IN} = 0$ in Fig.10, the voltage at the op amp's non-inverting input will be whatever voltage the capacitor has charged to. This will be amplified in the same way as any other input to give a DC output voltage.

If we assume the capacitor voltage starts at 0V at power-on, its DC voltage will steadily increase (positively or negatively). The output voltage will follow, proportionately with the gain. Eventually, the output voltage will reach the maximum voltage the op amp can output, which for simplicity we will assume is equal to the supply voltage.

At this point, the op amp will saturate and no longer function correctly as an amplifier. If an AC signal is present at the input, this will be superimposed on the DC voltage at the input and amplified. The AC part of the output will start to clip once the total voltage gets sufficiently close to the supply.

Providing bias current

The solution to the problem is straightforward: we connect a resistor (R_B) from ground to the non-inverting input to provide a path for the bias current, as shown in Fig.11.

A consequence of this is that C_1 and R_B form a high-pass RC filter with a -3dB cut-off $f_c = 1 \div 2\pi R_B C_1$ – a similar scenario to the inverting amplifier. The -3dB point is when the resistor and capacitor impedance magnitudes are equal in both cases.

Above the low-frequency cutoff, we would expect the capacitor to have a very low effective resistance, which we can simplify to assuming the capacitor is like a short circuit for AC signals at the frequencies of interest. With this assumption, we see that the input impedance of the circuit in Fig.11 is equal to R_B.

An advantage with the non-inverting circuit is that R_B is independent of the gain setting (via R_F and R_G) and can be a large value (100s of kilohms to megohms). Typically, R_B in the non-inverting circuit can be much larger than R_I in the inverting circuit, meaning that the circuit can have a high input impedance.

Consequently, much smaller capacitors can be used for the same low-frequency cut-off. This can reduce the cost and size of the circuit and/or allow better (more linear) capacitors to be used. Alternatively, if we need a specific input resistance, we can just set R_B to this value without worrying about interaction with the gain-setting.

Op amp bias currents are small (nanoamperes or picoamperes), so large R_B values do not prevent the op amp from obtaining sufficient bias or cause disruptive voltage drops (100pA through 100kΩ drops 10μV). However, any such voltage drop will be amplified and appear at the output.

It is worth noting that the same thing happens with the gain-setting resistors, so it is the balance between the amplified bias-current drops on both inputs that affects the output. In some circuits (particularly for DC), a resistor may be added to specifically balance the drops caused by bias currents.

The bias resistor will add some noise to the circuit – all resistors generate thermal or Johnson noise, which increases with both temperature and resistance. The larger the resistor, the more noise, so as always there are design trade-offs, in this case with noise and offsets vs input impedance and capacitor size.

Taking some time to fail

Returning to what happens if you fail to use a grounded resistor to supply bias, we can calculate the time taken for an amplifier to saturate due to the coupling capacitor charging. A fundamental equation for capacitors is $Q = CV$; the stored charge (Q) is equal to the capacitance times the voltage across the capacitor.

Current is the flow of change, so a constant current (I) flowing for time Δt onto a

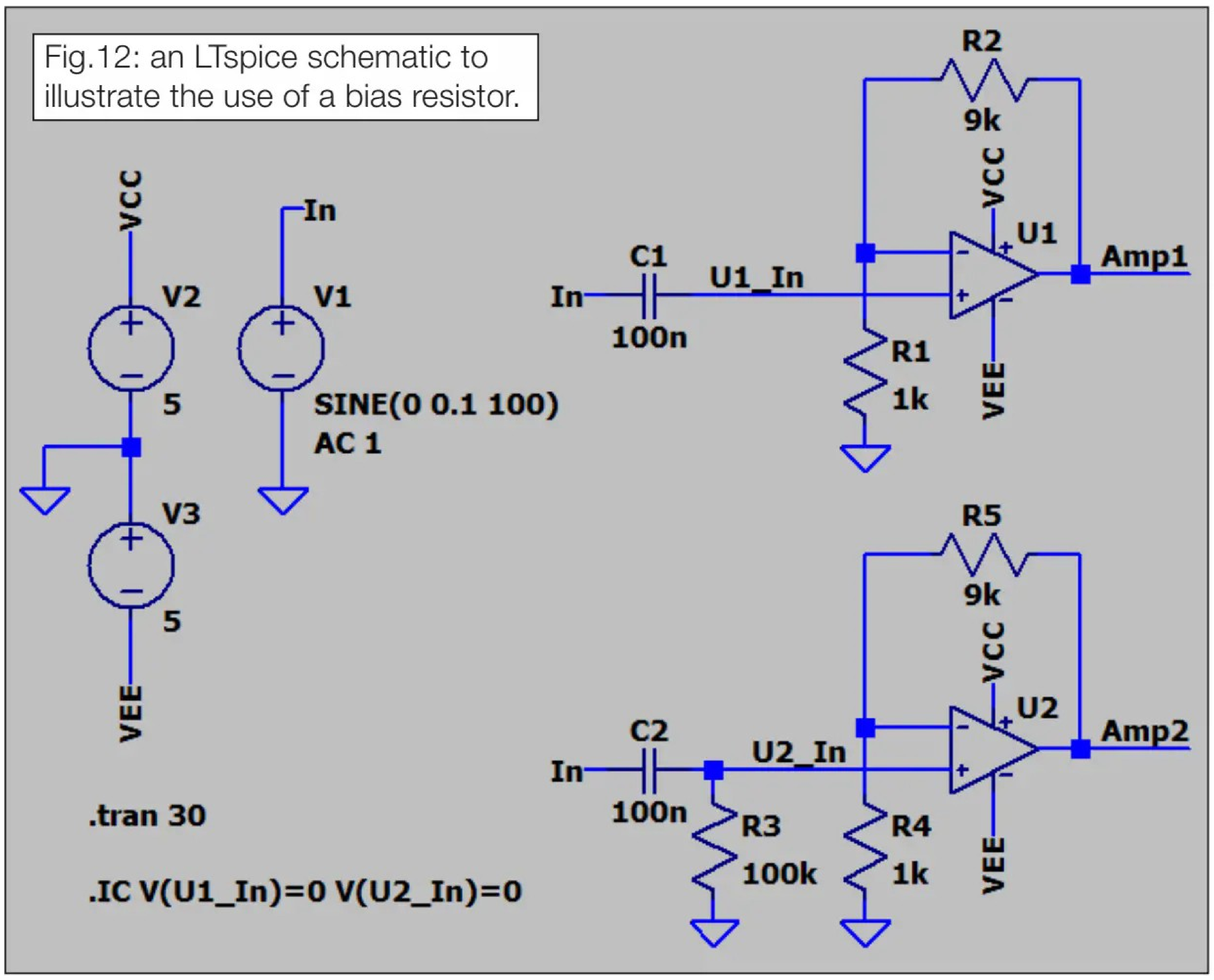


Fig.12: an LTspice schematic to illustrate the use of a bias resistor.

capacitor will add charge $\Delta Q = I\Delta t$ to that stored on the capacitor (Δ [delta] means ‘change in’). This will cause a voltage change of $\Delta V = (I/C)\Delta t$, or we can write the time taken for the voltage to change by ΔV as $\Delta t = (C/I)\Delta V$.

Consider a non-inverting amplifier (Fig.10) with a gain of 10, a bypass capacitor of 100nF, and an op amp with a 2nA bias current. If this was operating on a ±5V supply, it would need 0.5V on the capacitor to saturate (5V ÷ 10), which would occur about 25s after power-on, calculated as $(100 \times 10^{-9}\mathrm{F} \div 2 \times 10^{-9}\mathrm{A}) \times 0.5\mathrm{V}$.

Given that in other circuits, the capacitance could be much larger and the bias currents much lower than these figures, the time for the circuit to fail could run to minutes or even hours. This could be a major problem for the unwary, as a test of the circuit for a much shorter period may not reveal the problem.

Modern high-precision and bipolar low-bias op amps often have internal biasing circuits, which might seem to imply that the problem with bias currents charging the coupling capacitor would be avoided. However, external currents are not completely eliminated in these devices, so the same problem may arise, but take longer to manifest.

The bias current voltage-drop balancing resistors mentioned above are often not recommended for these op amps.

Simulating the problem

Fig.12 shows an LTspice simulation to demonstrate the effect of not using a bias resistor. The values are as in the example calculation above (gain=10, 2nA bias current). There are two copies of a non-inverting amplifier, with and without the bias resistor. The input is a 100Hz, 0.1V peak sinewave, which should produce a 1V peak output centred on 0V for the duration of the 30s simulation.

The op amp is modelled using LTspice’s UniversalOpAmp5 model, which allows you to set the input bias current for both inputs (right-click symbol to access the parameters).

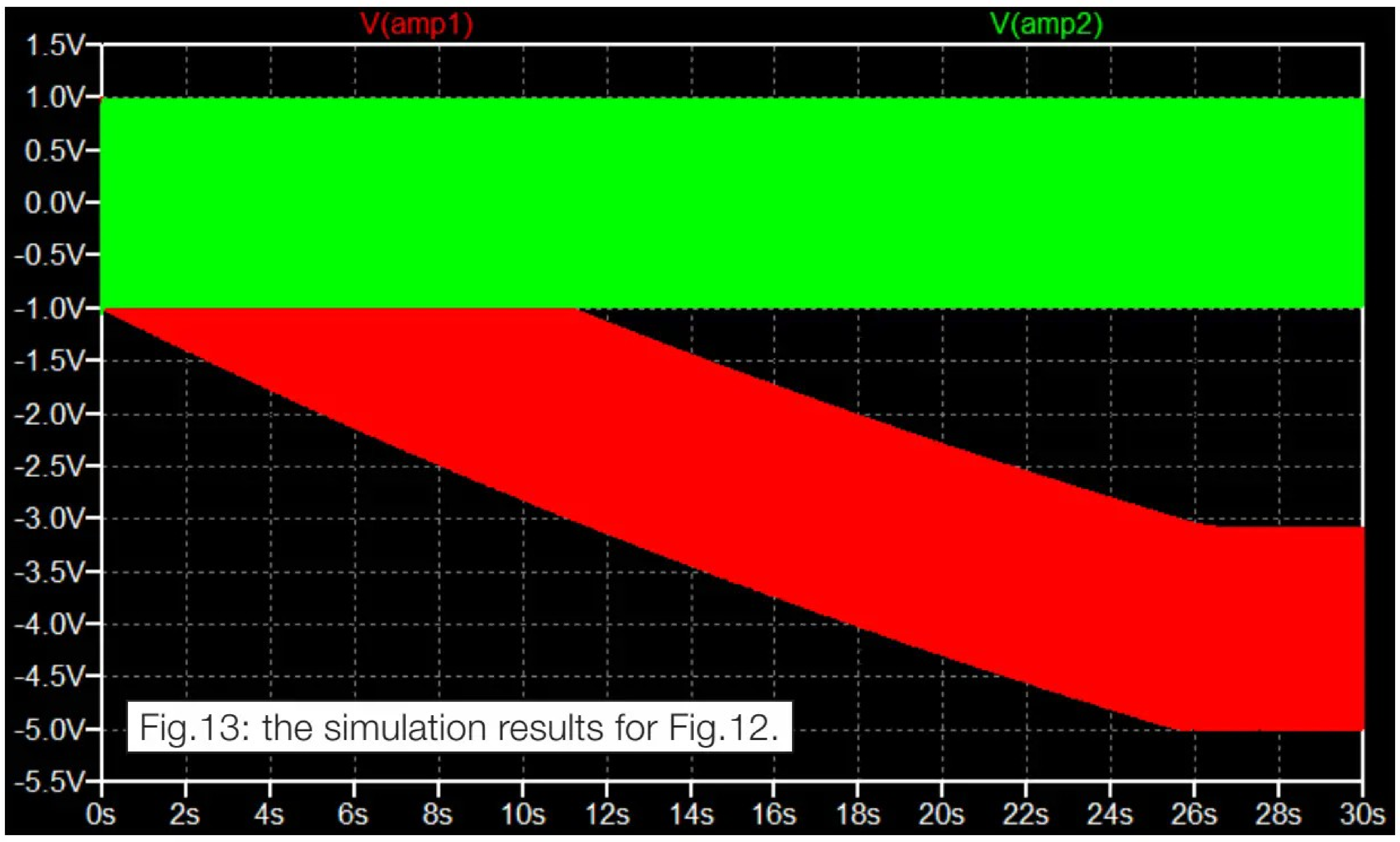


Fig.13: the simulation results for Fig.12.

Fig.13 shows the results of the full 30 second simulation. The circuit with the bias resistor maintains its output correctly centred on 0V throughout the simulation, whereas the one without the resistor drifts towards the negative supply, taking about 25s as calculated above.

Figs.14 & 15 show a few cycles of the signals near the start and end of the simulation. After 50ms (Fig.14), the signals are both centred on zero, although there is a slight shift down for the circuit without the resistor. At the end of the simulation (Fig.15), the output from the circuit with the bias resistor is unchanged, but the one without the resistor has drifted to a significant negative offset and the waveform is badly distorted.

The output does not completely saturate, as predicted by simply assuming the bias current is an unchanging ideal constant current source. The behaviour is, as might be expected, more complex, but the time estimate is sufficiently close given that this is a failing circuit.

Choice of op amp

There are a huge number of op amps available, and it can be daunting trying to choose one. Device manufacturers and component suppliers provide online interactive selection tables that can be very useful if you know what ranges of parameter values are appropriate.

Manufacturers often classify op amps according to application type or performance (eg, high precision, high speed, audio), which can help narrow the search.

Since the signal conditioning circuits we are considering here are operating on AC signals in the audio range and need to provide good signal integrity so as not to influence the measurements too much, devices listed by manufacturers for audio applications should be a good fit.

This classification should imply low noise and distortion characteristics, with sufficient bandwidth and slew rates for audio signals. We do not need very high precision DC operation or very high bandwidth into the GHz range.

Device packaging should be considered. I wanted to be able to use solderless breadboard to quickly construct and modify circuits as I was experimenting with REW. Therefore, I was looking for devices in DIP packages. This mainly means older devices as more recent ones are often only available in SMD packages.

This does not prevent their use with breadboards but means that you have to solder them to adaptors.

Power supply requirements are another consideration. I had a board providing ±5V supply available and preferred not to

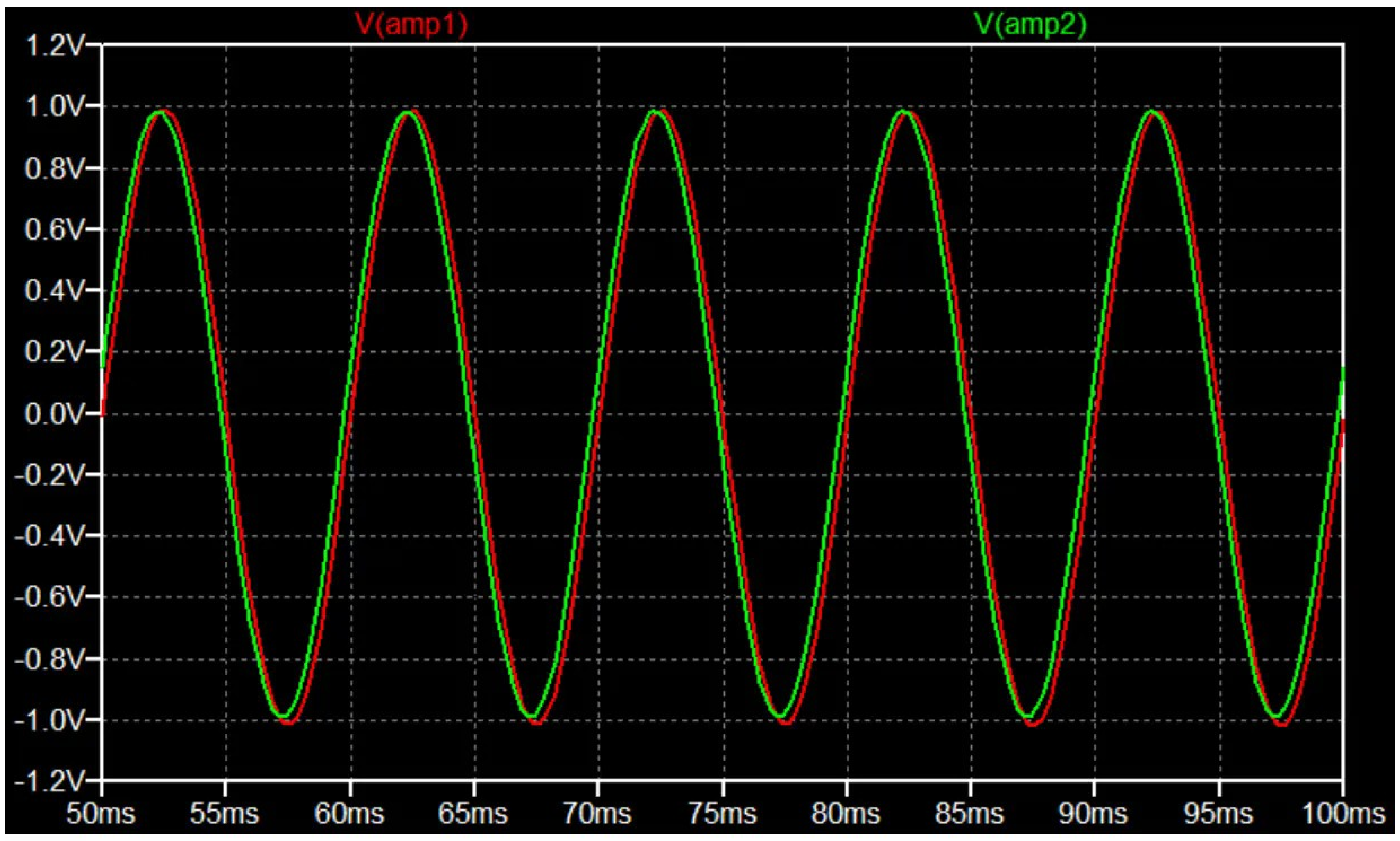


Fig.14: the initial simulated waveforms for Fig.12.

have the supply not go too far beyond the 3.3V used by the microcontroller implementing the filter. Different op amps have different minimum and maximum supply voltages, so this needs to be checked on the data sheet (and recall the earlier discussion about rail-to-rail types).

Op amps that you regularly see used in relevant applications may help guide your choice. For example, the NE5532 and NE5534 are widely used audio op amps that are available in DIP packages. However, here these devices would be operating at their minimum supply of ±5V, which may not give optimal performance. [Editor's note – our experience is that they still perform very well with a ±5V supply as long as the signals remain below about 2.2V RMS.]

The Texas Instruments page for the NE5534 suggested the OPA134 "high-performance audio op amp" as an alternative; this device can operate as low as ±2.5V and has other specifications that are suitable. Several other devices were also considered before selecting the OPA134.

When using unity gain buffers or other circuits with low gain, it is important to check the minimum circuit gain the op amp can be used with. Negative feedback provides significant advantages in compensating op amp imperfections and allowing us to accurately set gain just via resistor selection (we do not need to know the gain of the op amp itself).

The more negative feedback is applied, the easier it is for an unwanted phase shift (eg, due to stray capacitance on the input) to change the feedback to positive, causing instability (unwanted oscillation).

Low-gain op amps have the most feedback (percentage of signal fed back) so are more likely to be unstable. The gain of op amps is deliberately reduced at high frequencies to help overcome this, but that reduces the available bandwidth in higher-gain circuits. Therefore, some op amps are not designed to be stable in low-gain circuits, as this improves performance in higher-gain applications.

The NE5532 and OPA134 are unity-gain stable, but the NE5534 is stable for gains of three or more. A capacitor can be connected to the compensation pins of the NE5534 to facilitate use in lower gain circuits. Not all high-bandwidth op amps have this capability, though; check the data sheets.

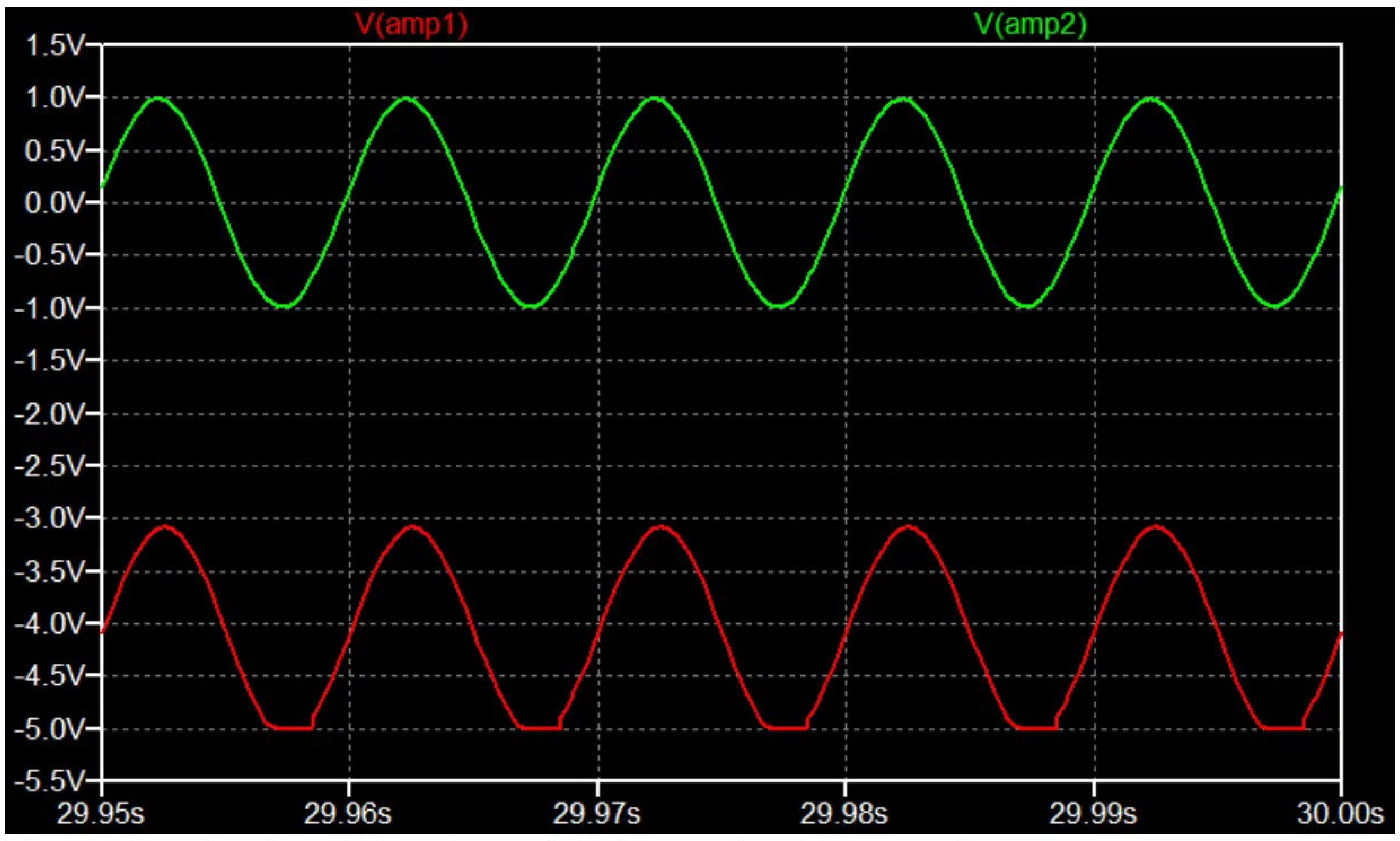


Fig.15: the simulated waveforms for Fig.12 after 30 seconds.

Noise, distortion and CMRR

The noise gain is higher than the circuit gain for inverting amplifiers, but the gains are the same for the non-inverting configuration. This means that, generally, the non-inverting configuration has lower noise. However, for AC-coupled circuits, the bias resistor is an additional source of noise in the non-inverting configuration that would not be present in a DC circuit.

Op amps are differential amplifiers, so ideally they ignore signals that are the same on both inputs (common-mode signals). Of course, they are not perfect, and the measure of their ability to do this is called the common-mode rejection ratio (CMRR). Op amps also have a maximum common-mode input voltage at which they will work correctly (specified as common-mode input range).

Although the inverting and non-inverting amplifiers built from the op amps are not differential, the common-mode characteristics of the op amp do impact circuit performance, particularly for the non-inverting configuration.

The virtual short circuit between the op amp's inputs in the non-inverting configuration means that a common-mode signal equal to the amplifier's input signal is applied to the op amp. This will have some influence on the output, so the effective gain will be different from that set by the designed resistor values (by a factor of 1/CMRR) and the signal distortion will also be slightly higher.

As most op amps have a high CMRR (eg, 70-100dB), the gain error is small, typically smaller than resistor tolerances (the error is about 0.03% for a 70dB CMRR). For the signal-conditioning circuit, the gain error does not matter, as the gain is manually adjusted to suit the signals.

The data sheet for the OPx134 (OP134, OP2134 & OP4134) discusses the fact that the input capacitance of the FET input stages used varies with the common-mode voltage, which increases distortion if the parallel combination of gain setting resistors is greater than 2kΩ, so adding a balancing resistance is recommended.

If the input signal amplitude of a non-inverting amplifier exceeds the maximum common-mode input voltage, problems that may occur including distortion, phase reversal (the output flips polarity), saturation, a significant increase in input current and even damage to the device.

The common-mode input range for the OPx134 is to within 2.5V of the supply rails, so the signal amplitude should be less than 2.5V peak on ±5V supplies. For the digital filter setup, the largest

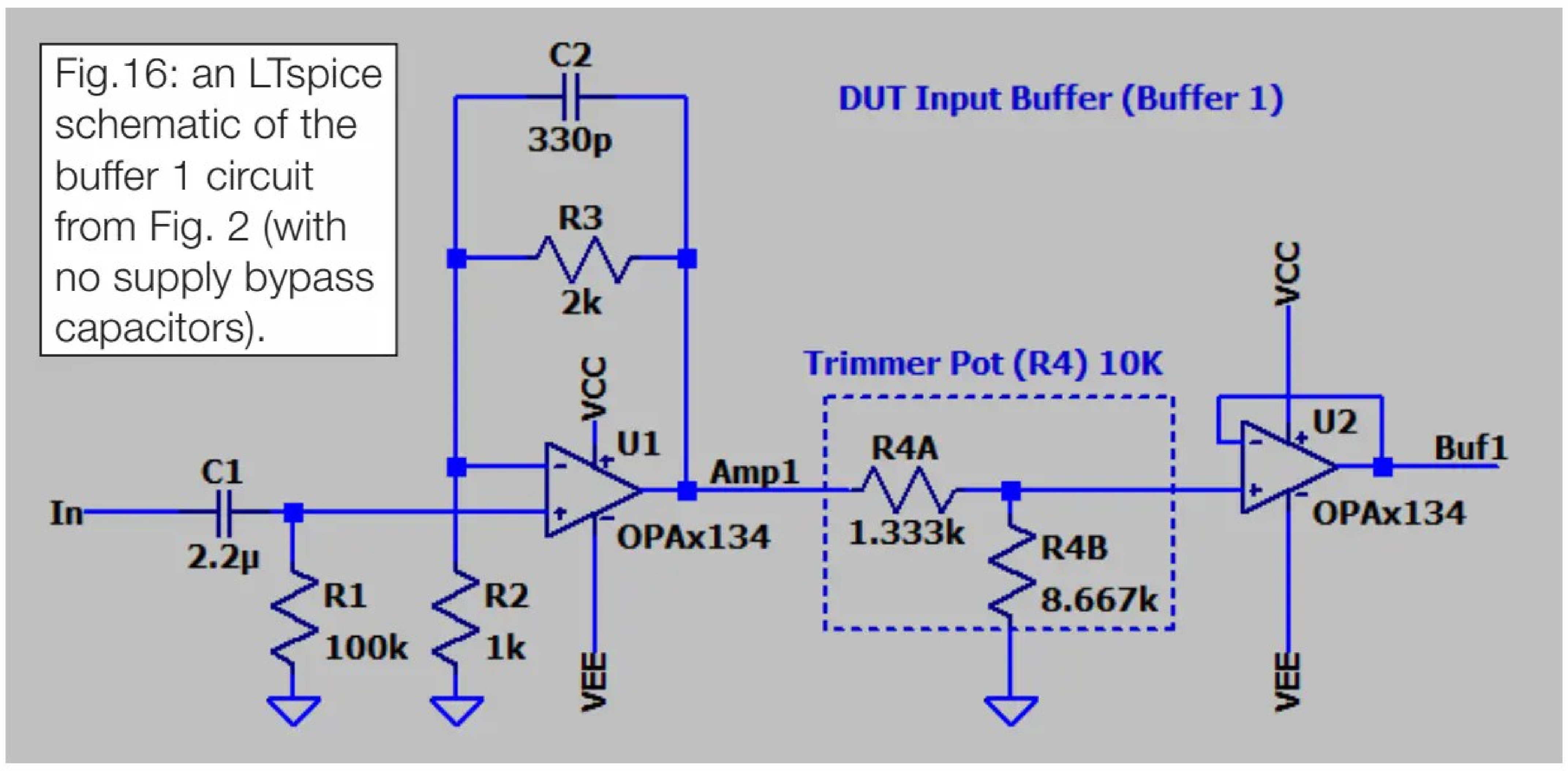


Fig.16: an LTspice schematic of the buffer 1 circuit from Fig. 2 (with no supply bypass capacitors).

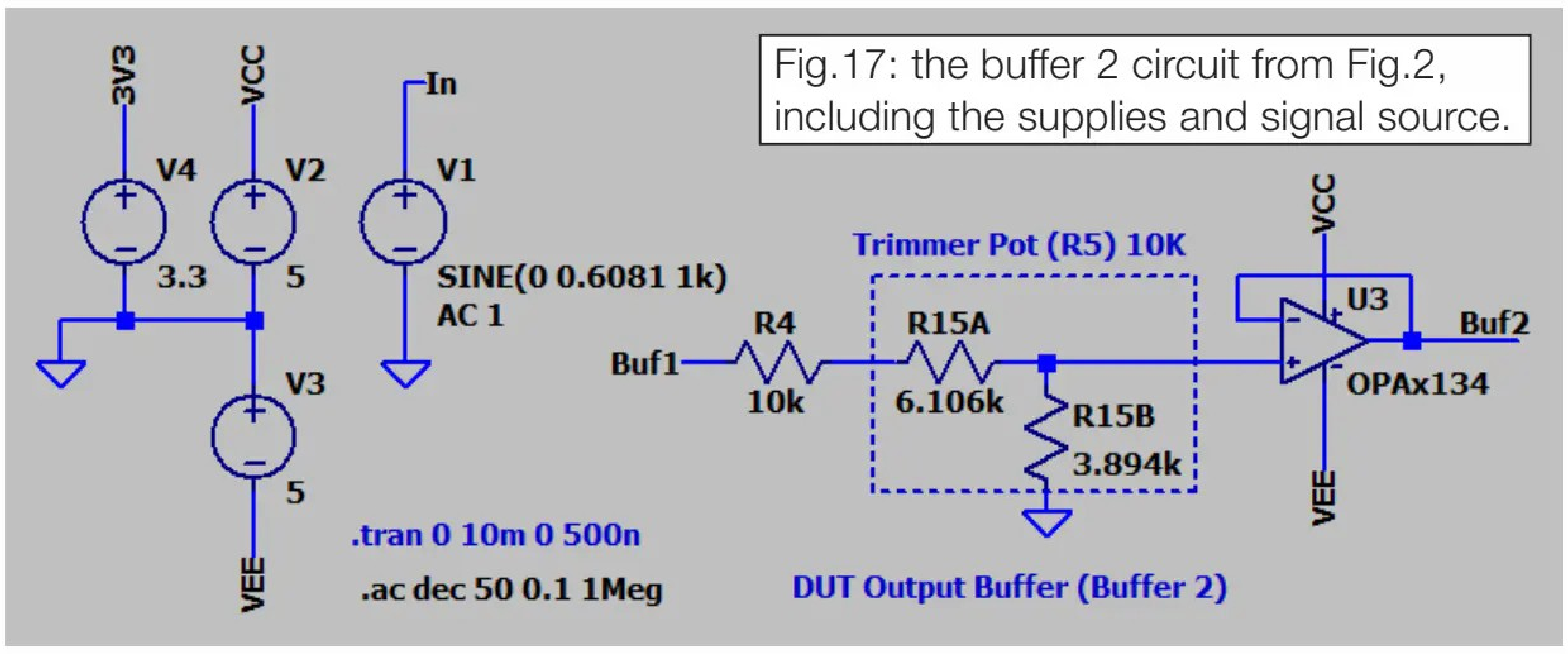


Fig.17: the buffer 2 circuit from Fig.2, including the supplies and signal source.

signal is 1.65V peak, so well within the required range. To handle larger signals, a higher supply voltage can be used, or a rail-to-rail op amp.

Like most modern op amps, the OPx134 has protection against phase reversal, even for several volts beyond the specified common mode range.

Buffer circuits and simulation

LTspice simulation schematics for the buffer 1 and buffer 2 circuits from Fig.2 are shown in Figs.16 & 17. This is one simulation file in which the output of buffer 1 is connected to buffer 2 (there is no model of the device under test, as this is just a simulation of the buffers). Fig.17 also includes the power supplies, signal source and simulation commands.

The supply bypass capacitors (Fig.6) included in the practical circuit (100µF electrolytic for both supplies on the breadboard and pairs of 100nF ceramics for each chip) are not included because the simulation supplies are ideal voltage sources. The model of the OPx134 op amp was downloaded from Texas Instruments and added to LTspice.

Buffer 1 includes a non-inverting amplifier (like Fig.11) with a gain of three times (+9.5 dB), a bias resistor of 100kΩ and a 2.2µF coupling capacitor, giving a low-frequency cutoff of 0.72Hz, well below the 20Hz minimum frequency we were measuring for the digital filter.

As an aside, electrolytic capacitors work well for AC-coupling as long as the -3dB point is kept sufficiently low (usually no more than a few Hz). They are somewhat nonlinear, but this arrangement keeps the distortion due to non-linearities outside of the audible range.

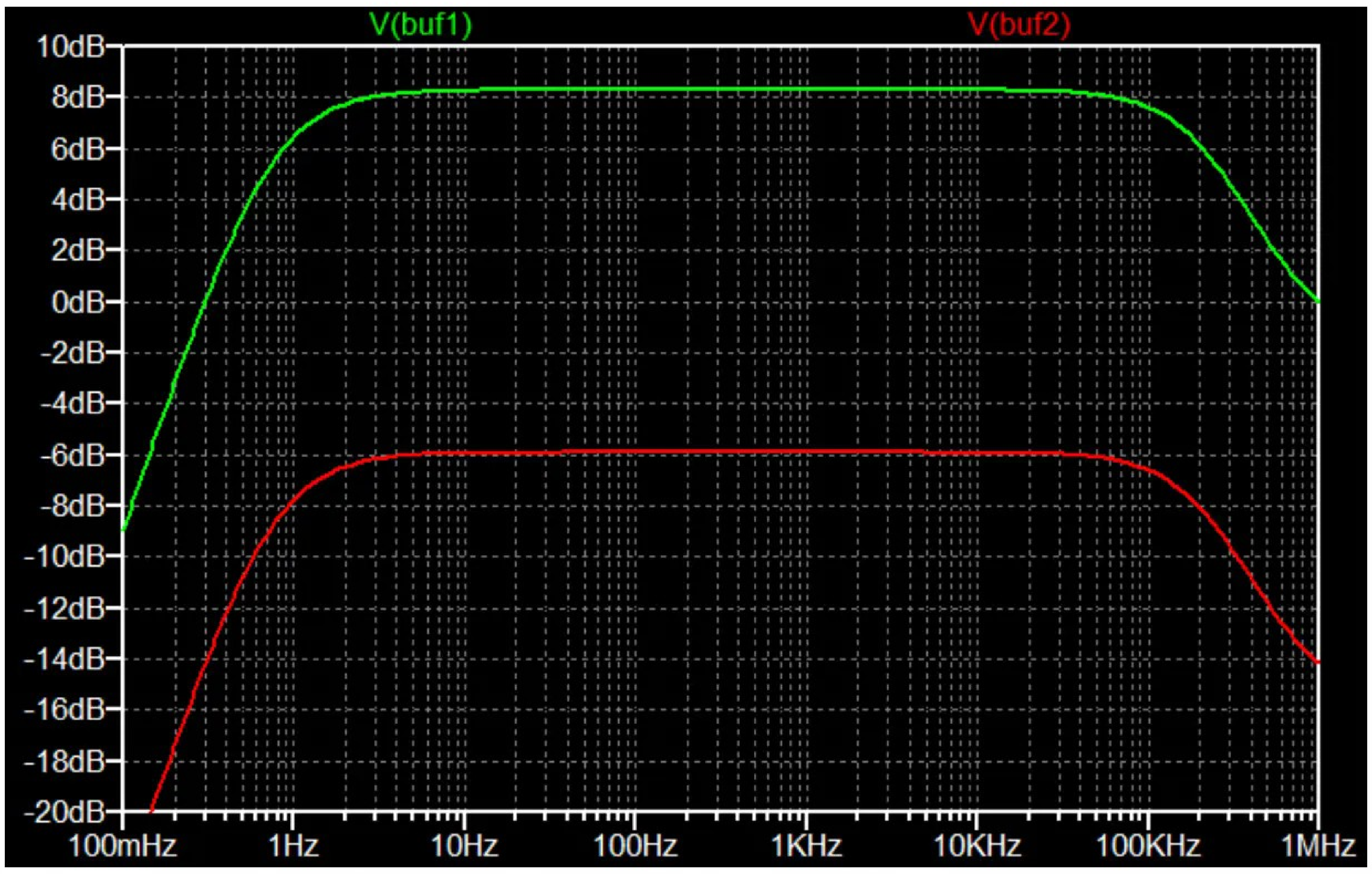


Fig.18: the results of simulating the circuits in Figs.16 & 17.

The amplifier is followed by a variable attenuator (like Fig.8). The practical circuit uses a trimmer potentiometer, but the simulation has fixed resistors for simplicity, with values selected for an overall gain of 2.6× (+8.3 dB), as discussed last month and recapped above.

A 100kΩ trimmer could have been used in place of R_1 to save an op amp, but for experimenting with these circuits, it was helpful to have one operation per op amp, and I didn't have a suitable trimpot to hand!

A capacitor added across the feedback resistor reduces the gain to unity (1) at high frequencies. The -3dB point (buffer 1 gain 5.3dB) is at around 255kHz. Feedback capacitors are commonly used to improve amplifier stability.

The gain reduction will help if there are unwanted high frequencies in the input. More usually, an RC low-pass filter and/or ferrite bead would be included in the input circuit to address this if needed.

Buffer 2 is simply another attenuator, again with fixed resistors to model the trimmer and set the attenuation as required (to 0.195 or -14.2dB). This gives an overall gain for the two buffers of 2.6 × 0.195 = 0.507 (8.3 – 14.3 = -5.9dB). With the gain of the line driver (not included here) being about 1.97× (+5.9dB), the overall gain for the circuit in Fig.2 is unity (0dB), as required.

10kΩ resistor R4 provides finer control of the gain in the 0-0.5× range than just using a trimmer. The input is from buffer 1, so an input op amp (as shown in Fig.8) is not needed. The output op amp is needed, as the line driver has a low input impedance.

The simulation results (AC Analysis) in Fig.18 confirm that the circuits have the required gain values and are flat over the 20Hz to 20kHz range required to measure the digital filter frequency response.

When the key interest in a simulation is AC analysis to confirm gain and frequency response, it is worth running a transient simulation to check that there are not unexpected problems with the waveforms. It is not impossible to make mistakes with the schematic drawing or design that will not be obvious in an AC analysis. The schematic in Fig.17 can be configured for a sinewave simulation.

Next month

Next month, we will continue to look at aspects of the circuit design theory and practice for the signal conditioning circuitry used with the frequency response measurement of the example digital filter.

Project by Stephen Denholm

This straightforward piece of test equipment measures capacitor values over a wide range, from about 10pF to 10,000µF (10mF). It's easy to assemble with all through-hole parts, fits into a UB1 Jiffy box, and won't break the bank either.

Digital Capacitance Meter

Digital Capacitance Meter
Capacitor Cx Value
= 4372.9 uF
Press MEASURE button
to continue ....
Hi
Mid
Lo
RANGE
On/Off
Cx
MEASURE
Hi: 100µF ~ 10,000µF
Mid: 1µF ~ 100µF
Lo: ~10pF ~ 1,000nF
Denholm Electronics
50 V 4700µF

I occasionally need to measure values of large electrolytic capacitors (up to at least 6800µF) but have been restrained by the limited capacitance measurement ranges of the DMMs I have. To overcome this, I've resorted to setting up a test circuit using a digital oscilloscope to measure a capacitor's value by measuring their charge time.

This worked well but was time consuming. I explored Silicon Chip magazines looking for a relatively simple capacitance meter project that I could expand my development skills on and build. I found quite a few articles on the subject, ranging from very simple to quite advanced designs.

Of particular interest was the Circuit Notebook item "PIC capacitance meter measures charging time" by William Andrew (July 2008; siliconchip.au/Article/1874). It was a little too basic for my requirements, but I liked the relatively simple design concept, which appeared to work.

I therefore decided to develop a similar design that was also PIC-based, would use the charging time measurement concept, was relatively simple to build and compact, covered a range from about 10pF up to about 10,000µF, and was powered by a standard 9V battery.

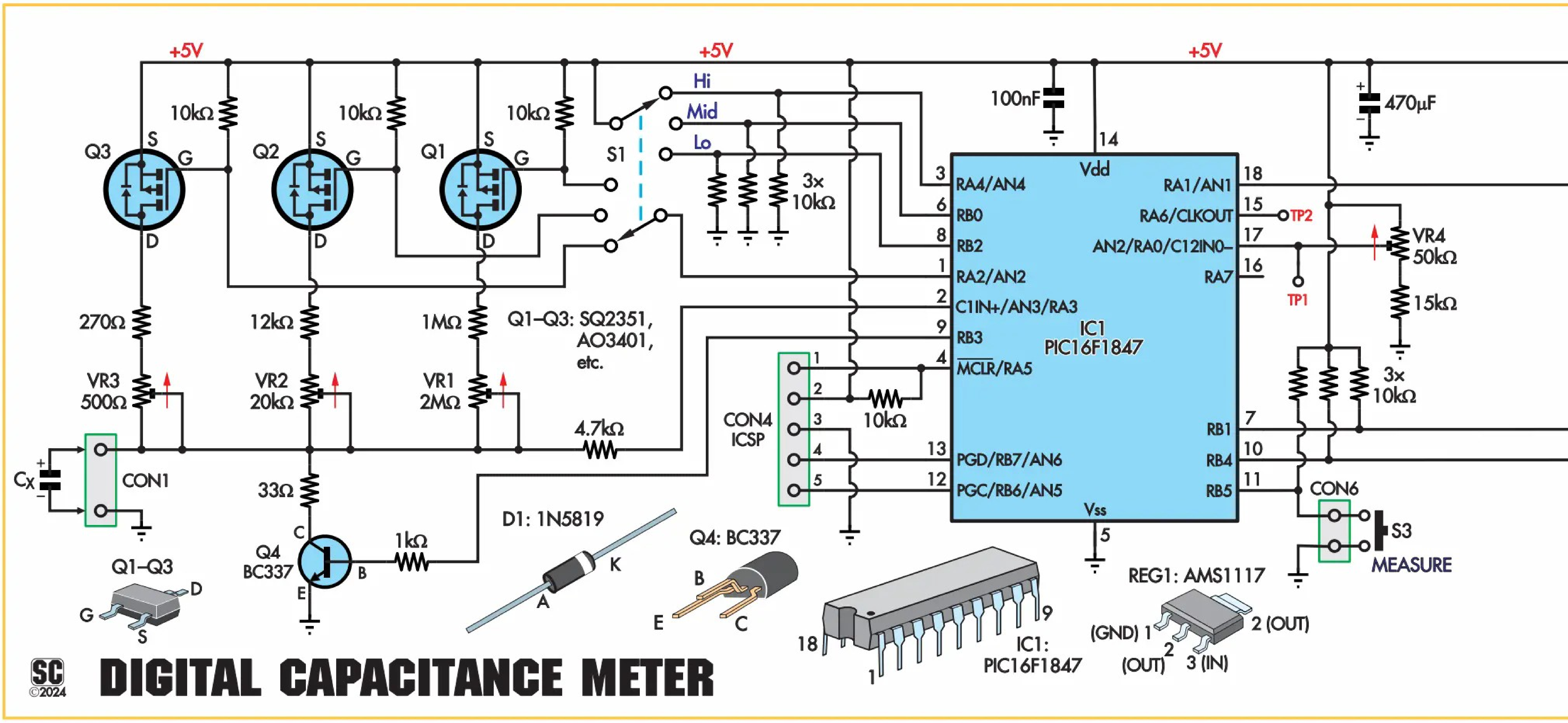
I was also inspired by Jim Rowe's article on low cost 1.3-inch OLED displays in the September 2023 issue of *Practical Electronics*. I thought I would have a go at also incorporating one of those low-power display modules into my design.

Circuit details

As shown in Fig.1, my circuit uses an 8-bit enhanced mid-range PIC16F1847 microcontroller unit (MCU). It has three capacitance ranges selected by switch S1 and shows the measured value of the capacitor under test (C_X) on the 1.3-inch OLED display (MOD1). The OLED is also used to display any over/under range or battery voltage warning messages that are necessary.

The measurement operating sequence is commenced by pressing pushbutton switch S3. The MCU will then first ensure that C_X is fully discharged by switching on Q4 for a short period, then off, discharging it via the 33Ω resistor. It then starts charging capacitor C_X via one of Mosfets Q1, Q2 or Q3 and the associated series resistance.

At the same time, it starts the MCU's 16-bit Timer1, which operates with a counting interval of 1µs. The charging


+5V
+5V
+5V
Q3
Q2
Q1
10kΩ
10kΩ
10kΩ
S
G
D
Hi
Mid
Lo
S1
3x 10kΩ
100nF
470µF
14
Vdd
3 RA4/AN4
6 RB0
8 RB2
1 RA2/AN2
2 C1IN+/AN3/RA3
9 RB3
4 MCLR/RA5
13 PGD/RB7/AN6
12 PGC/RB6/AN5
RA1/AN1 18
RA6/CLKOUT 15
AN2/RA0/C12IN0– 17
RA7 16
RB1 7
RB4 10
RB5 11
Vss 5
IC1 PIC16F1847
TP1
TP2
VR4 50kΩ
15kΩ
3x 10kΩ
270Ω
12kΩ
1MΩ
Q1–Q3: SQ2351, AO3401, etc.
VR3 500Ω
VR2 20kΩ
VR1 2MΩ
4.7kΩ
CON4 ICSP
10kΩ
CX
CON1
33Ω
Q4 BC337
C
B
E
1kΩ
D1: 1N5819
K
A
Q4: BC337
CON6
S3
MEASURE
REG1: AMS1117
Q1–Q3
18
9
1
IC1: PIC16F1847
(GND) 1
2 (OUT)
2 (OUT)
3 (IN)
SC ©2024

DIGITAL CAPACITANCE METER

voltage developed across C_X is then measured by the MCU's Comparator1 positive input (C1IN+, pin 2) and compared to the voltage applied on its negative input C12IN0– (pin 17).

As soon as the charging voltage exceeds the voltage at C12IN0–, the comparator stops Timer1, initiates a program interrupt and passes control back to the main program, where the Timer1 count register values are used to calculate the capacitance.

As the source voltage for charging C_X is the 5V V_{DD} supply, the comparator C12IN0– input is set to 63.2% of V_{DD}, nominally 3.16V. This ensures that the comparator operation and hence measurement time will always be equivalent to one RC time constant of the capacitor under test.

That simplifies the calculation to C_X = Timer1 count (µs) ÷ selected range series resistance, scaled accordingly. For the Lo, Mid and Hi capacitance ranges, the MCU calculations use series resistance values of 2MΩ, 25kΩ or 500Ω, respectively. It also means that, even if the output of the 5V regulator drifts with temperature or time, the measurements should remain accurate.

The actual values used in the circuit are provided by the fixed/variable resistance combinations VR1 + 1MΩ, VR2 + 12kΩ and VR3 + 270Ω, which are switched into or out of circuit by the MCU via Mosfets Q1, Q2 and Q3. I used P-channel SMD devices as, particularly for the Hi range, they need low on-resistances to slightly improve the measurement accuracy.

Suitable PNP transistors such as BC858s with base resistors of say 1kΩ to 3.9kΩ may work reasonably well, but with a small reduction in measurement performance. However, I have not tried that arrangement.

Performance

Performance-wise, my meter has been providing quite accurate and repeatable results across all three ranges. I have confirmed this occasionally by checking the meter's range extremities against the calibration capacitors that I now keep for such a purpose. On the Lo range, it is necessary to keep the meter leads short to minimise any stray capacitance.

In the MCU program code, I have allowed compensation for zero-offset in the Lo range calculations, which significantly improves the capacitance measurements for values below 1nF and surprisingly allows the meter to achieve quite accurate and consistent results down to about 10pF.

This zero-offset value compensates for some inherent MCU program instruction cycle time, which starts to dominate the measurements for very short capacitance charging durations. It also compensates for the stray capacitance inherent in the physical construction of the meter and the short leads I use.

The zero-offset value is hard-coded, but it is not too difficult to change by re-compiling the code and uploading it to the MCU if necessary. I did briefly think about adding an auto-zeroing function to the meter design but decided it wasn't worth the extra effort for my particular requirements, especially if I always stick to using the same meter leads.

Construction

The board, coded 04111241 and measuring 80 × 100mm, is a double-sided design, but there are only a few top-layer tracks that can easily be replaced by wire links, as you will see in the photo of my prototype. So if you are etching the board yourself, start by fitting the four wire links you can see in that photo; they are also visible as top-layer tracks in the overlay diagram, Fig.2.

Also note that there are four SMD components that mount on the underside: Mosfets Q1-Q3 and regulator REG1. They are shown in 'X-ray' fashion in Fig.2. Start by soldering them in place while the board will still fit flat on your bench.

Q1-Q3 are all the same types and REG1 is in a different package, so it should be obvious which goes where and in what orientation. Do make sure that the leads are sitting flat on the board before soldering and not sticking up in the air, which would indicate that the part is upside-down.

Tack each part by one pin and check that all the leads are over the matching PCB pads. If not, remelt that joint

Digital Capacitance Meter

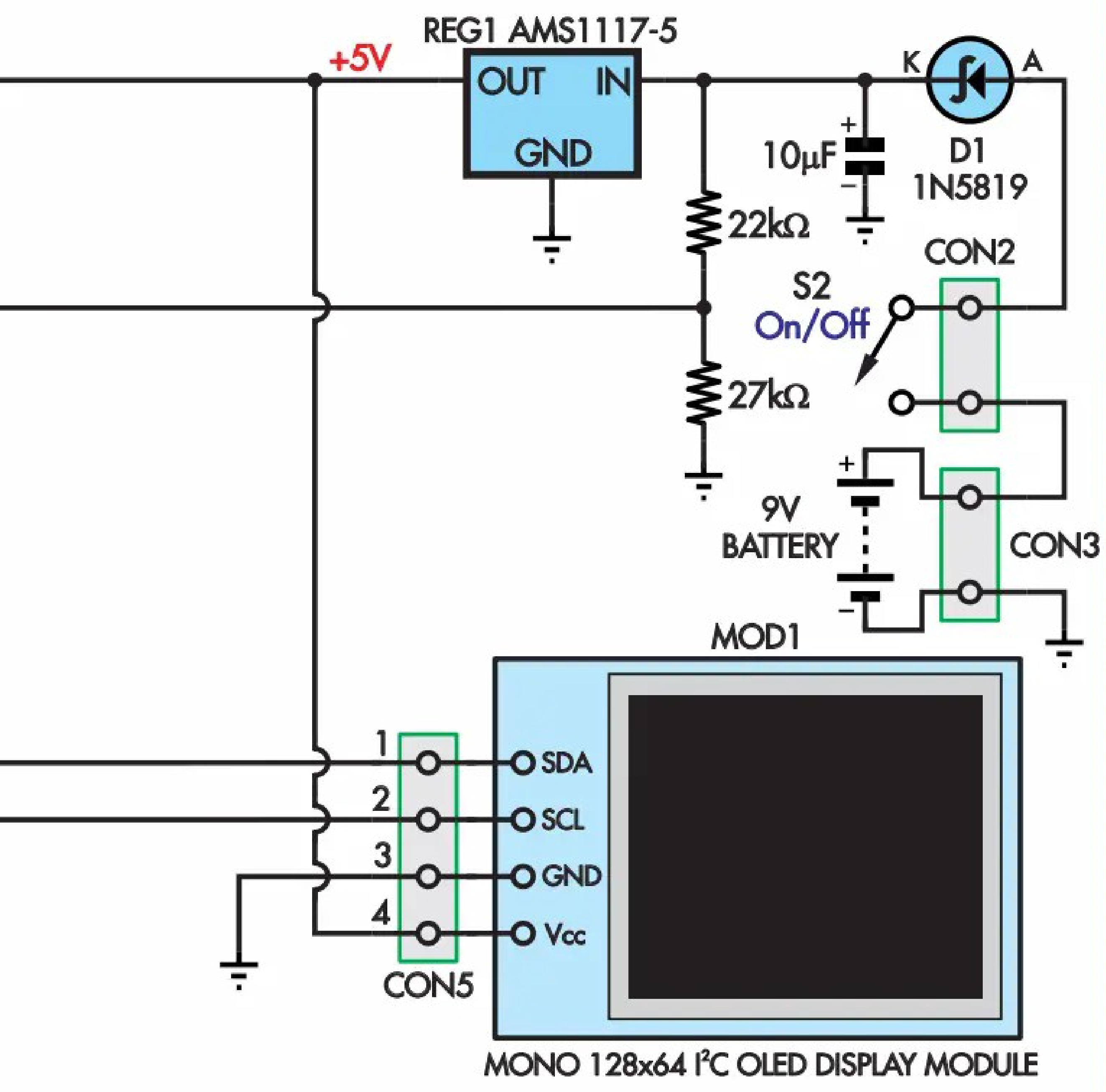
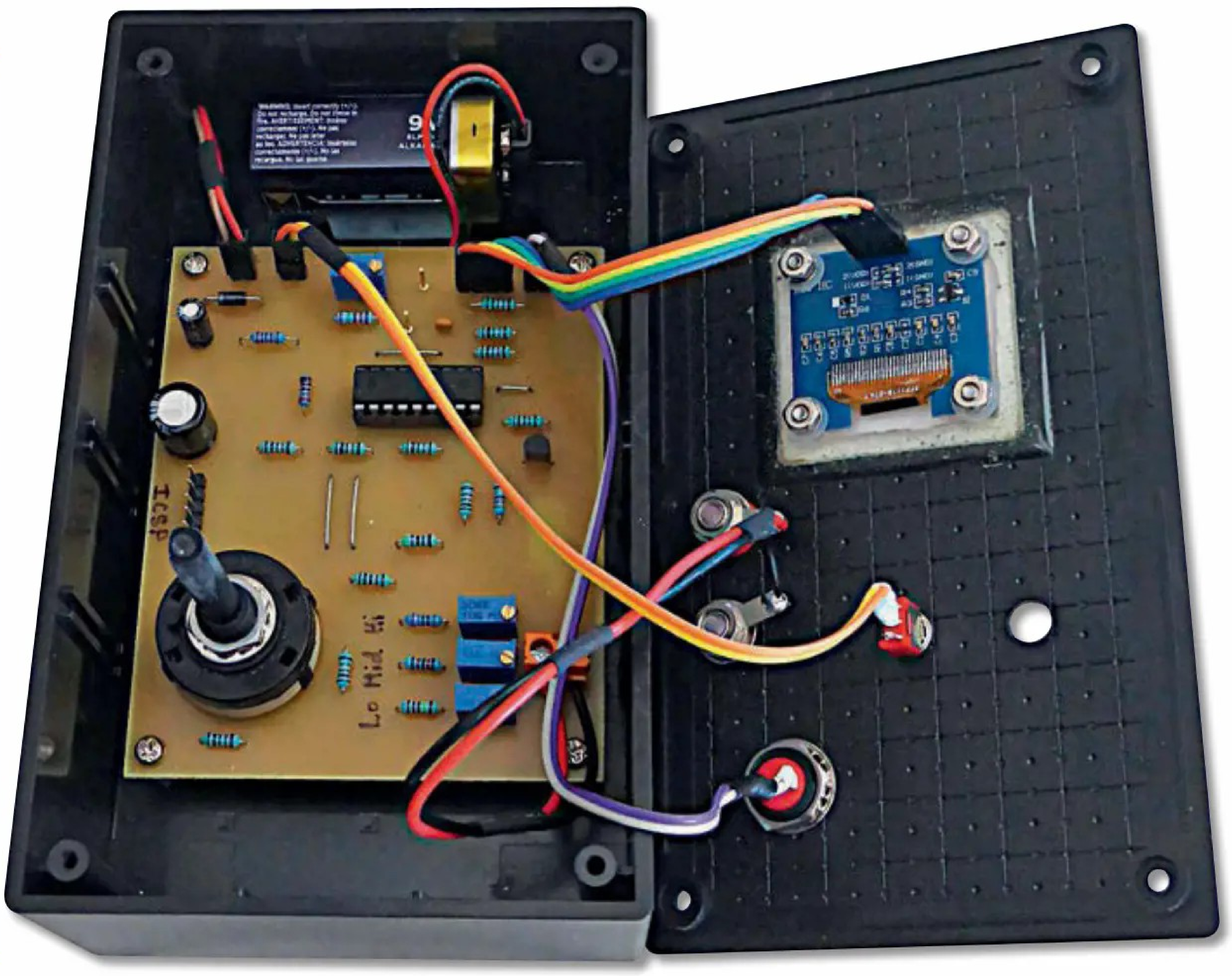

+5V
REG1 AMS1117-5
OUT
IN
GND
K
A
D1
1N5819
10µF
+
22kΩ
CON2
S2
On/Off
27kΩ
9V
BATTERY
CON3
MOD1
1
2
3
4
SDA
SCL
GND
Vcc
CON5
MONO 128x64 I²C OLED DISPLAY MODULE

Fig.1: the circuit diagram for the Capacitance Meter. S1 is used to switch the capacitance range.



and gently nudge it into place. Once it's properly aligned, solder the remaining leads and then refresh the first joint.

Next, flip the board over and solder all the resistors in place. They are mounted with the leads bent quite close to the bodies. Follow the overlay diagram to see which values go where. There is just one diode, so fit that now, making sure its cathode stripe goes towards the top edge of the board as shown in Fig.2.

You don't have to use a socket for IC1, but it makes it easier to swap that chip if that ever becomes necessary. Solder either the socket or IC1 directly to the board, but in either case, make sure it is orientated with its notched (pin 1) end towards the top of the PCB.

Solder terminal block CON1 in place now. We recommend that its wire entry holes are kept towards the left-hand side, although you can insert the wires from either end.

Next, fit the headers (CON2-CON6), 100nF capacitor (which is not polarised) and transistor Q4 (orientated as shown). Note that CON4 is only required if you plan on (re)programming IC1 in-circuit. You could leave the other headers off and solder wires directly to the board, but we suggest using headers to make assembly (and if required later, disassembly) much easier.

Mount the four trimpots next, making sure the adjustment screws all go towards the bottom of the board as per

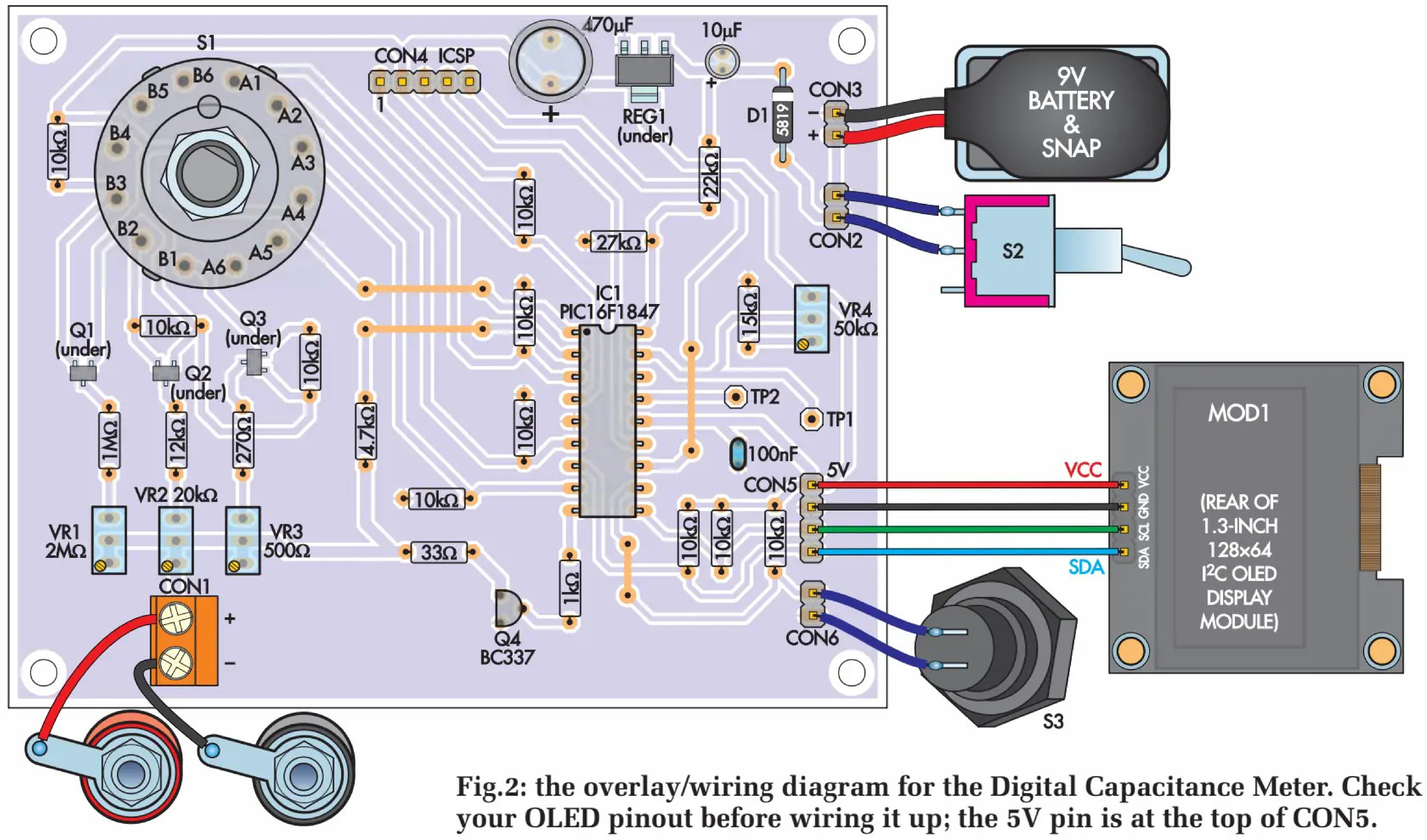

S1
B6
A1
B5
A2
B4
A3
B3
A4
B2
A5
B1
A6
10kΩ
CON4 ICSP
1
470µF
+
REG1
(under)
10µF
+
D1
5819
CON3
−
+
9V
BATTERY
&
SNAP
22kΩ
10kΩ
27kΩ
CON2
S2
Q1
(under)
10kΩ
Q3
(under)
Q2
(under)
10kΩ
IC1
PIC16F1847
10kΩ
15kΩ
VR4
50kΩ
TP2
TP1
1MΩ
12kΩ
270Ω
4.7kΩ
10kΩ
100nF
5V
CON5
VCC
SDA
SDA SCL GND VCC
MOD1
(REAR OF
1.3-INCH
128x64
I²C OLED
DISPLAY
MODULE)
VR2 20kΩ
VR1
2MΩ
VR3
500Ω
10kΩ
33Ω
10kΩ
10kΩ
10kΩ
1kΩ
CON1
+
−
Q4
BC337
CON6
S3

Fig.2: the overlay/wiring diagram for the Digital Capacitance Meter. Check your OLED pinout before wiring it up; the 5V pin is at the top of CON5.

Fig.2. They are all different values, so don't get them mixed up.

Now solder the two electrolytic capacitors in place, ensuring that the longer (positive) lead goes into the bottom hole in each case. The negative striped ends of the cans should be near the top edge of the PCB.

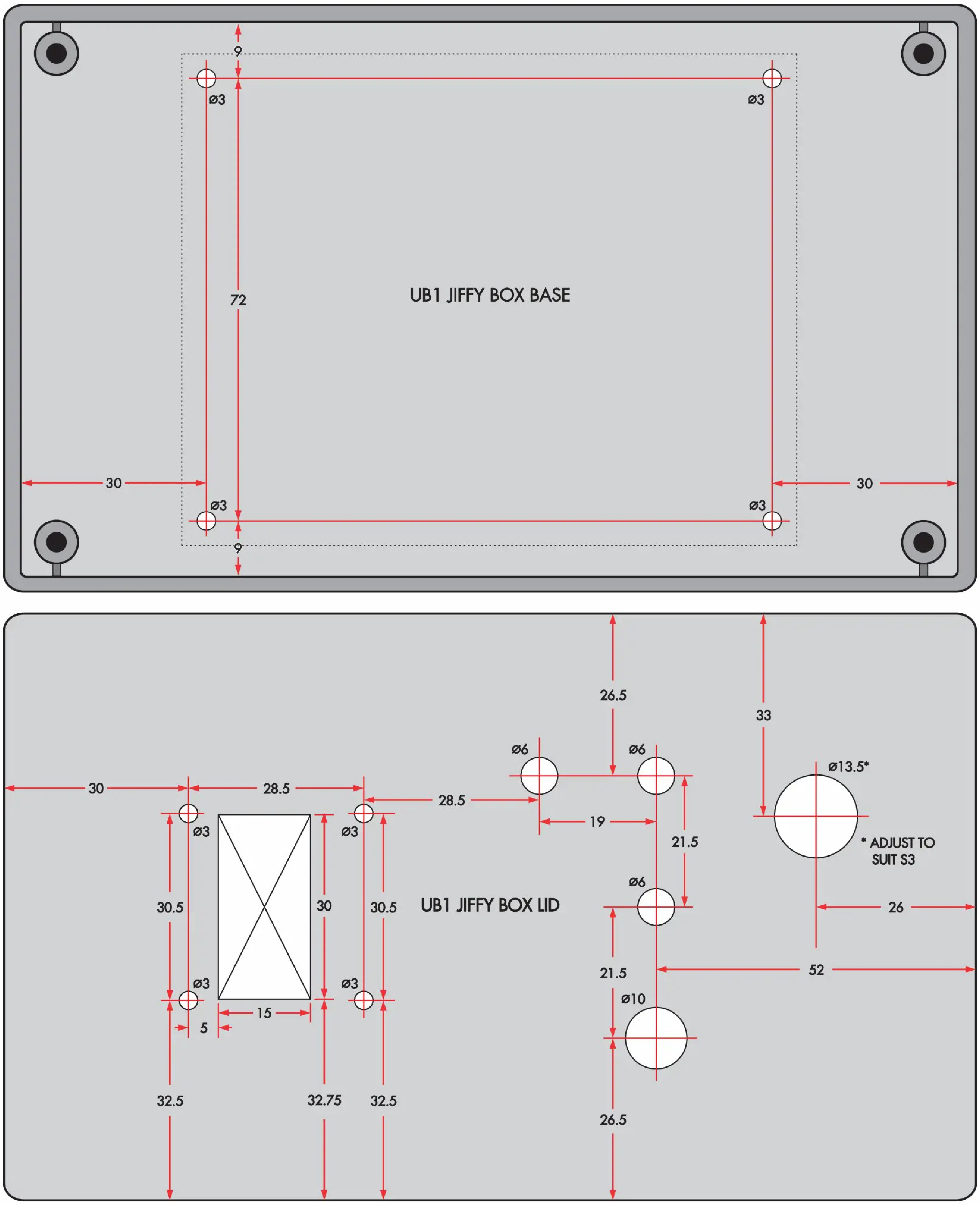
PCB pins for test points TP1 and TP2 are not strictly required if you have a double-sided board, as you can simply insert DMM probes into the plated through-holes. If you have a single-sided board, you will need to solder PCB pins into the two test point holes.

Rotary switch

The last part to mount directly to the PCB is the rotary switch. It is a two-pole type. As supplied, it will probably have six positions, but we only need three. To change that, undo the nut and remove the washer from the shaft. Prise up the stop washer and rotate the switch fully anti-clockwise, then re-insert the stop washer with its pin going into the second hole between the moulded "3" and "4".

Check that it now only switches through three possible positions. If not, change the position of the stop washer and try again. Once it's correct, put the lock washer back over the shaft and tighten the nut on top.

In my build, the switch shaft length as supplied was just long enough to pass



Figs.3 & 4: the cutting diagrams for the base and lid of the Jiffy box. You have some flexibility with the locations cutouts on the lid, as they're mounted off the board.

All diagrams are shown at actual size, and all dimensions are in millimetres.

through the front panel with enough poking through to attach the knob. The exact length required depends on the height of the spacers used to mount the PCB in the box and the knob you're using.

Ideally, you should temporarily mount the PCB in the box so you can check how much to cut off (if any). To do that, you will first need to drill PCB mounting holes in the base of the box and at least one hole in the lid (for the rotary switch shaft). The PCB mounting hole positions are shown in Fig.3 and the lid holes in Fig.4.

With the shaft cut to length, remove the PCB from the box and solder the switch to it. There are two possible orientations, so match the switch to the photos and overlay.

The next job is to mount the remaining parts on the front panel/lid and solder wires with female DuPont headers ready to plug into the headers on the PCB. If you haven't already, finish making the holes in the lid as per Fig.4, after reading the next two paragraphs.

Regarding the OLED screen, you can see from the photos that I used countersunk head screws, Nylon washers and nuts to mount it to a clear acrylic sub-panel, then glued that panel to the inside of the lid using epoxy. I did it this way as the acrylic panel provides some protection for the OLED screen; the screw heads are hidden under the front panel label.

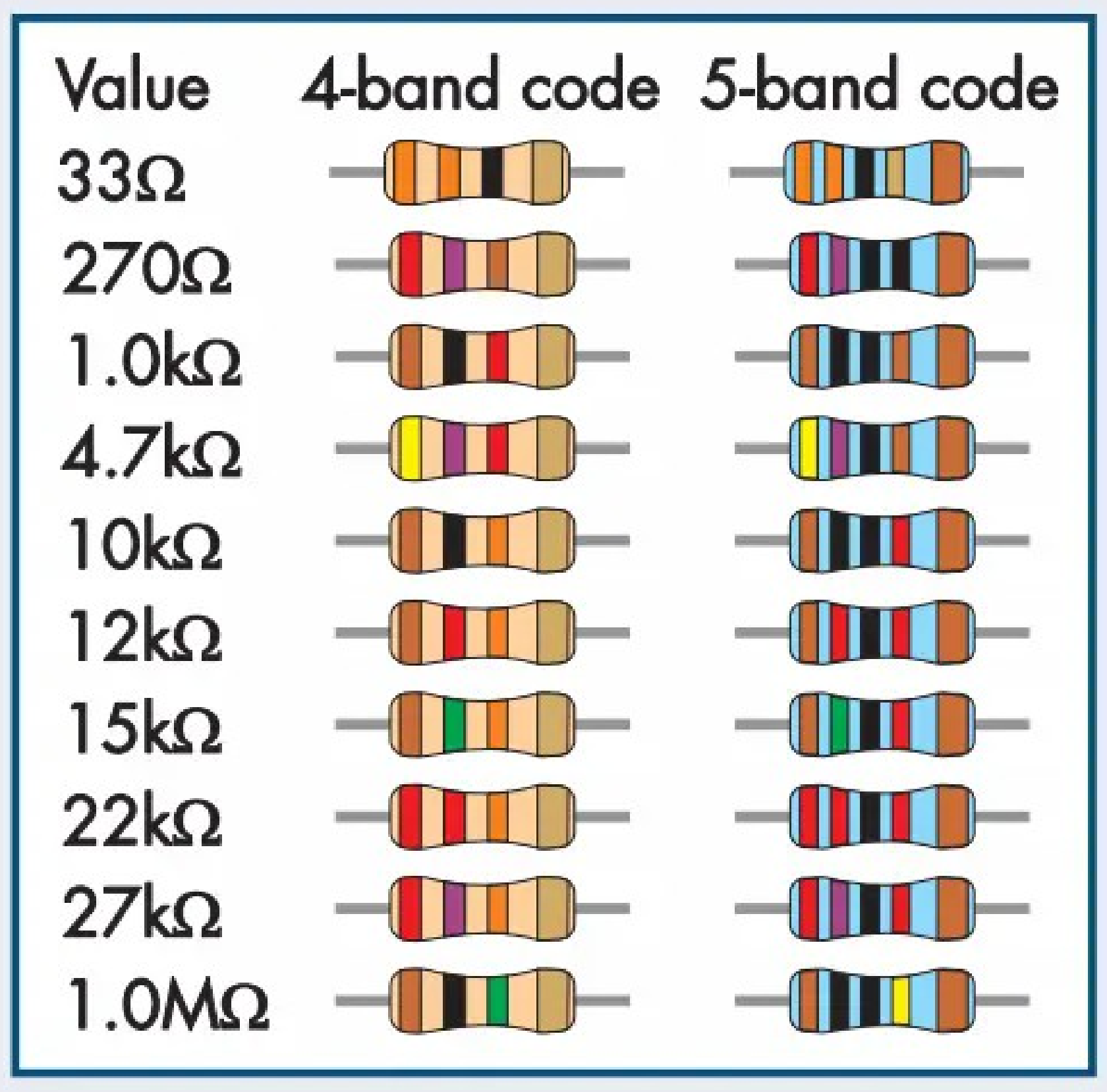
You could use the same approach, or mount the OLED directly to the lid using the holes shown in Fig.4. However, if you do that, note that even if you countersink the holes on the outside, the screws will probably still project above the surface of the lid due to its thinness. You may be able to cover them with a label but it's better to use my approach, if possible, if you want a flat panel label.

If you use my approach, use washers to space the OLED screen from the acrylic panel so the screen isn't crushed when you tighen the screws.

Strip off pairs of DuPont jumper wires from the ribbon for the 9V battery snap and switches S2 & S3. Strip off a set of four for the OLED. Cut them so that you have bare wires on one end, then solder them to the panel-mounting parts (check the OLED pinout with reference to Fig.2). For the two banana sockets, use medium-duty hookup wire (or similar) in two different colours instead.

Parts List – Digital Capacitance Meter

1 single- or double-sided PCB coded 04111241, 80 × 100mm
1 UB1 Jiffy box **[eg, Bud Industries CU-1945, Farnell 4903638]**
1 panel label, 100 × 160mm
1 1.3-inch (33mm) 128×64 pixel I²C OLED display module (MOD1)
  [Silicon Chip SC5026 or SC6511, AliExpress 32847652911]
1 3mm clear acrylic sheet of approximately 43 × 41mm
  (for mounting the OLED module)
1 2-pole 6-position sealed PCB-mounting rotary switch (S1)
1 miniature panel-mount SPST toggle switch (S2)
1 panel-mount momentary NO pushbutton switch (S3)
1 small-to-medium knob to suit S1
1 2-way 5.08mm pitch terminal block (CON1)
3 2-pin headers, 2.54mm pitch (CON2, CON3, CON6)
1 5-pin header, 2.54mm pitch (CON4; **optional**, for ICSP)
1 4-pin header, 2.54mm pitch (CON5)
1 red panel-mount binding banana socket
1 black panel-mount binding banana socket
1 pair of banana plug to crocodile clip test leads
1 2MΩ top-adjust multi-turn trimpot (VR1)
1 20kΩ top-adjust multi-turn trimpot (VR2)
1 500Ω top-adjust multi-turn trimpot (VR3)
1 50kΩ top-adjust multi-turn trimpot (VR4)
1 18-pin DIL IC socket (**optional**)
1 9V battery snap
1 9V battery retaining clip
1 9V battery
5 M3 × 6mm panhead machine screw
8 M3 × 6mm countersunk machine screw
4 M3 × 10mm tapped spacers
4 Nylon M3 washers
5 M3 hex nuts
10 short (~100mm) female/female DuPont jumper leads, joined in a ribbon
2 100mm lengths of medium-duty hookup wire (red & black)
1 100mm length of 1.5mm diameter black/clear/white heatshrink tubing
2 PCB stakes/pins (**optional**)



Value	4-band code	5-band code
33Ω		
270Ω		
1.0kΩ		
4.7kΩ		
10kΩ		
12kΩ		
15kΩ		
22kΩ		
27kΩ		
1.0MΩ		

Semiconductors
1 PIC16F1847-I/P 8-bit microcontroller programmed with 0411124A.HEX,
  DIP-18 (IC1)
1 AMS1117-5.0 or similar 5V 1A LDO linear regulator, SOT-223 (REG1)
3 AO3401(A) or SQ2351ES P-channel logic-level Mosfets, SOT-23 (Q1-Q3)
1 BC337 45V 800mA NPN transistor, TO-92 (Q4)
1 1N5819 40V 1A schottky diode (D1)

Capacitors
1 470µF 10V radial electrolytic
1 100µF 10V ±5% tantalum **[Vishay Sprague 293D107X5010D2TE3]**
1 10µF 50V radial electrolytic
1 2.2µF 50V ±5% MKT **[TDK B32529D0225J000]**
1 100nF 50V ceramic or multi-layer ceramic
1 100nF 63/100V ±5% MKT **[Vishay BFC237012104]**

Resistors *(all ¼W 1% axial)*
1 1MΩ
1 27kΩ
1 22kΩ
1 15kΩ
1 12kΩ
10 10kΩ
1 4.7kΩ
1 1kΩ
1 270Ω
1 33Ω

You can then plug everything into the headers on the PCB, using Fig.2 as a reference, and screw the two banana socket wires into the terminals of CON1. Ensure the wire routing is correct for the 9V battery, OLED screen and wires to CON1.

With IC1 out of its socket, switch on power and check the voltage between pins 5 and 14 of that socket. You should get a reading between 4.5V and 5.5V. If not, switch off and check for faults.

Assuming it's close to 5V, switch off and insert IC1 in its socket, ensuring it has the correct orientation and that none of the leads fold up under the body when you do so.

If IC1 has not been programmed, you can now power the device back on and connect an in-circuit programmer to CON4, with its pin 1 marking to the left as shown. Use software like Microchip's free MPLAB IPE to load the HEX file, which you can download from siliconchip.au/Shop/6/532

You can then switch it back on and check that the screen display comes up normally. If so, you can proceed with calibration. Otherwise, power it off and check your soldering and parts placement.

Calibration

To initially calibrate the meter, set the voltage at test point TP1 (IC1's negative comparator input voltage) to 3.16V by adjusting trimpot VR4. There is no ground test point; you could use negative (bottom) terminal of CON1.

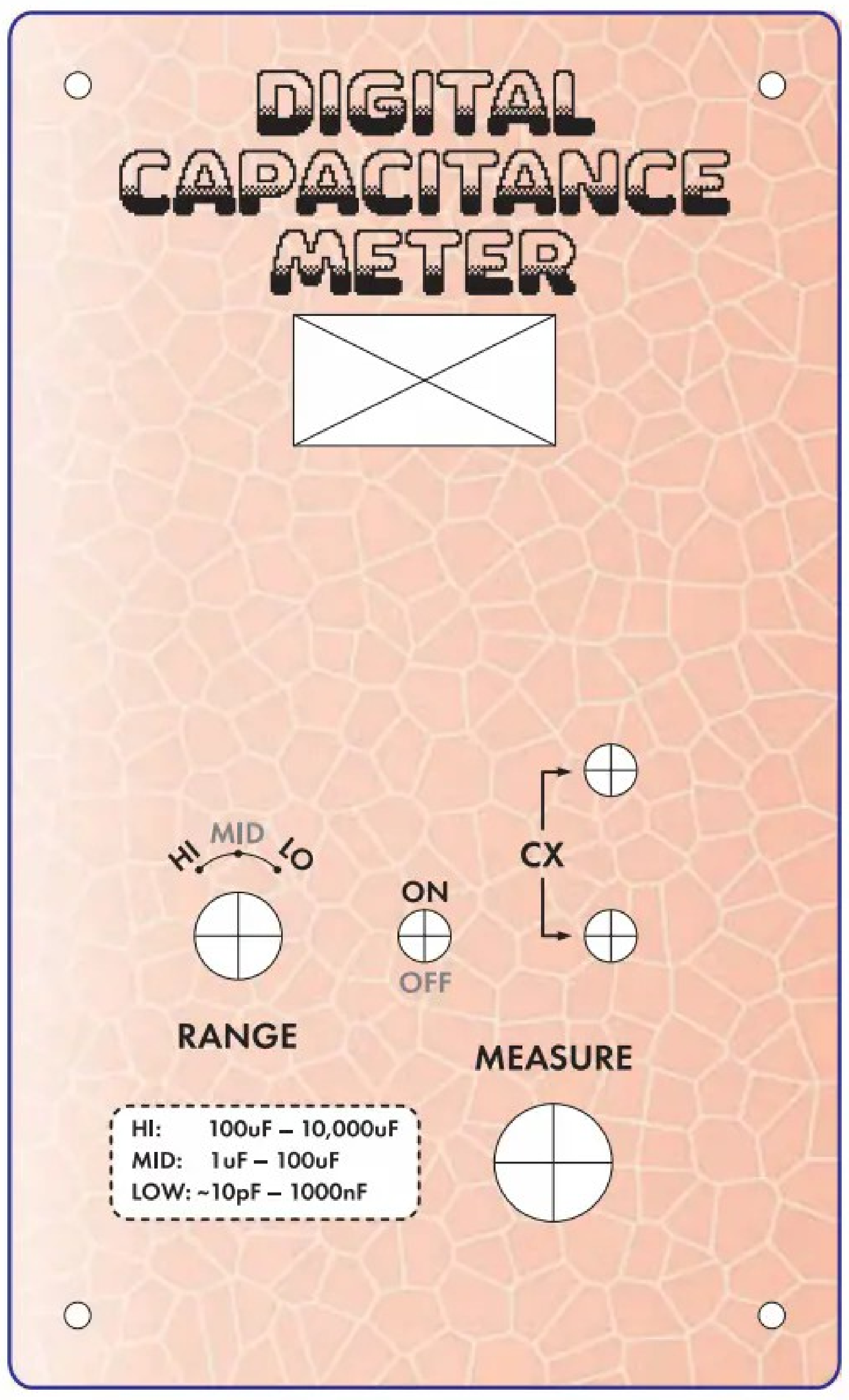
Next, for each range in turn, make repeated capacitance measurements of a calibration capacitor of known value while adjusting the selected range trimpot (VR1-VR3) to progressively obtain a calibrated value very close to the known capacitances.

The parts list includes suggestions of three low-cost 5% tolerance capacitors that could be used, although sourcing the larger values may not be easy (DigiKey and Mouser have suitable parts).

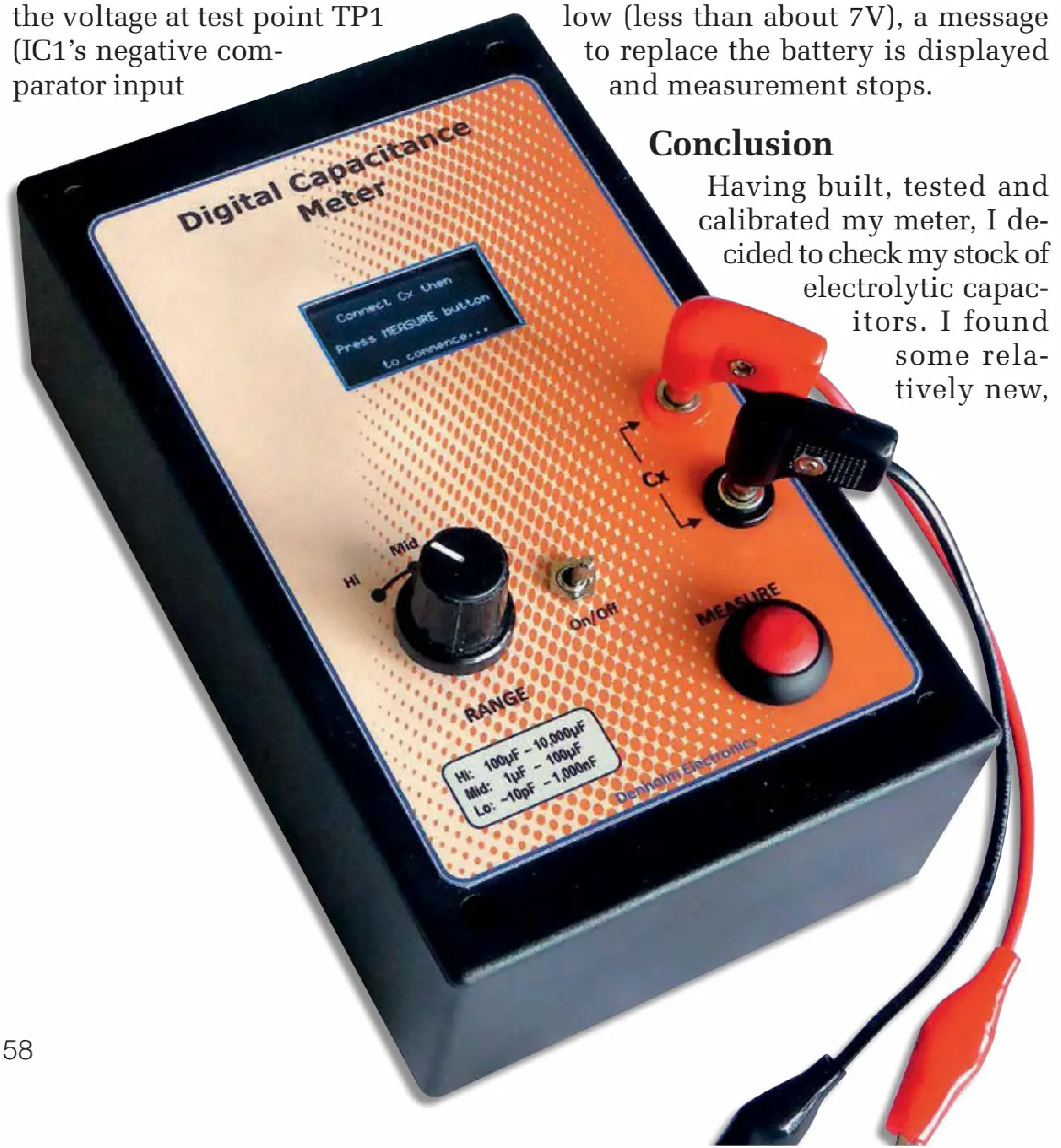
Cycle through the ranges and adjust each to get the correct measurement until you are only making minimal adjustments.

In operation, once the measurement and calculation of the capacitance is completed, the MCU displays the value on the OLED in units of either pF, nF or µF depending on the range selected and size of the capacitor under test. If the measured value is out of range, a warning is shown to select a higher or lower range if possible.

Also, before any measurement of C_X commences, the MCU checks the battery voltage and a warning message appears if it is low. If the voltage is too low (less than about 7V), a message to replace the battery is displayed and measurement stops.



The photo below is of the finished Digital Capacitance Meter with crocodile clips attached. Our version of the front panel label (shown above at 50% actual size) is available to download from siliconchip.com.au/Shop/11/585



Conclusion

Having built, tested and calibrated my meter, I decided to check my stock of electrolytic capacitors. I found some relatively new, unused electrolytic capacitors with values nowhere near their labelled value and not within the specified tolerance. In fact, I would say these capacitors had been incorrectly labelled or manufactured, as they were that far out!

This was rather concerning as these components had been sourced from reputable suppliers. Buyer beware, as they say!

I built the Touchscreen Wide-Range RCL Box from the June & July 2021 issues of *PE* a few years ago now. I've found it to be a very handy device. When I first built it, I thoroughly checked all the resistance values and found these to be well within the ±1% tolerance, which was great. However, I did not check the C and L values.

So, out of interest, I decided to do a quick check on the capacitance values with my new meter. Surprisingly, I found two capacitors well outside (>30%) the ±10% tolerance I was expecting, even though I'm sure I had purchased SMD capacitors with specified tolerances of ±10% or better. I also performed a check with a DMM on capacitance range and got very similar results.

I'm now waiting on a rainy day to do some further diagnostics on the RCL box. PE

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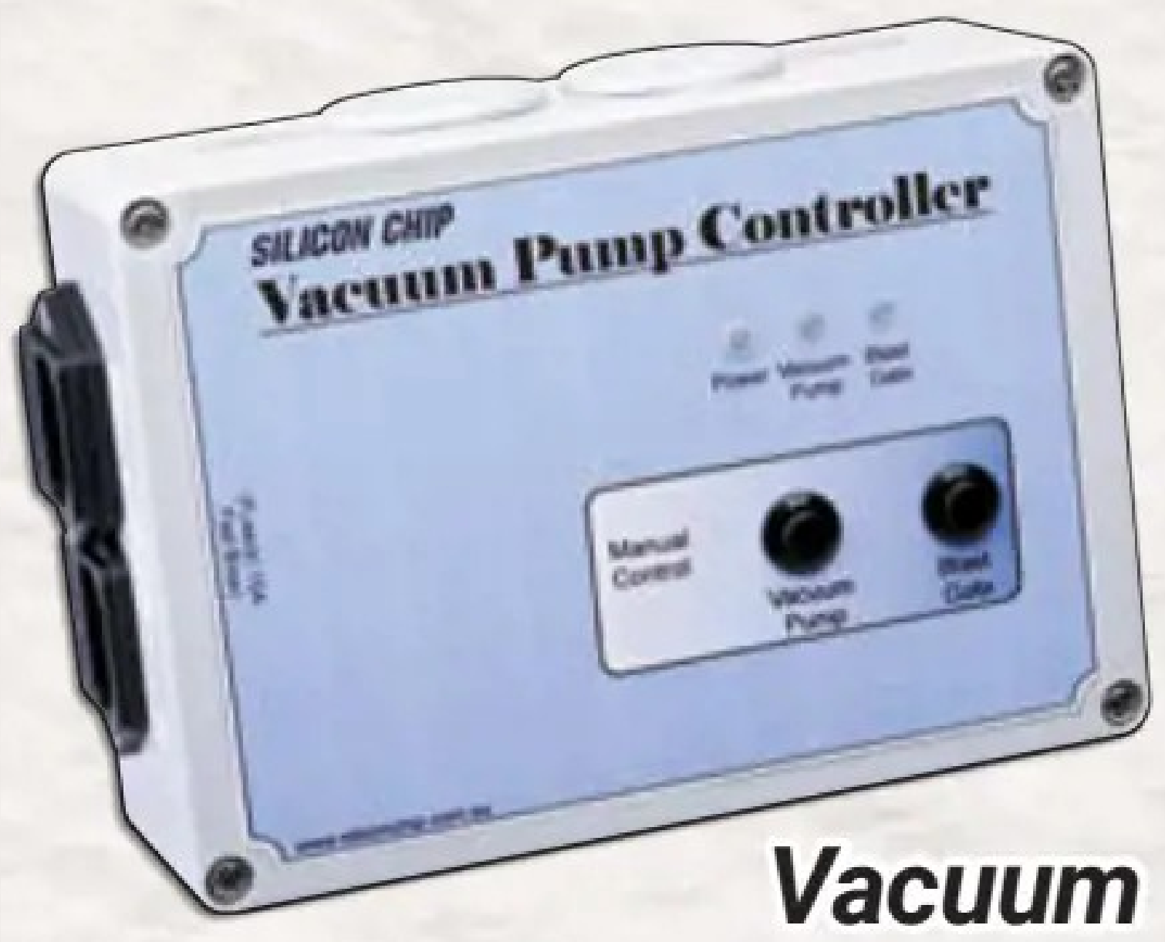
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Net Work

Alan Winstanley

This month's column brings updates on the SpaceX Starship program, welcome news for Windows 10 users, some puzzlement over using smart battery chargers and the end of an era for one of the most popular names in electronics book publishing, Bernard Babani.

It's easy to forget that it will soon be 60 years since people first set foot on the moon. NASA's Apollo 11 carried Neil Armstrong and Buzz Aldrin in a lunar module that landed on July 20th, 1969, with a global audience watching the mission unfold live on their television screens.

A YouTube video uploaded by the Associated Press captures the tension of the time at https://youtu.be/hzApsIPHRwo

The idea of using hydrogen and liquid oxygen (LOX) as propellants for a rocket is attributed to Russian scientist Konstantin Tsiolkovsky (1837-1935). Tsiolkovsky developed the earliest principles of aerodynamics and liquid-propellant rockets, and his pioneering work was published at the turn of the last century.

Ideas for a Sputnik-like spacecraft were scribbled out nearly 150 years ago in 1883, as shown in Wikipedia at https://en.wikipedia.org/wiki/Konstantin_Tsiolkovsky

Today, the development of Starship, the world's largest rocket, is continuing unabated. Starship could carry a payload of 150-250 metric tonnes, and SpaceX has an ambition to land on Mars. The eleventh test flight was scheduled to launch in mid-October, just before this magazine went to press.

SpaceX also planned to test a new configuration of engine burns during the final descent phase, before the Super Heavy booster plunges into the Gulf of Mexico.

Once in orbit, the upper stage would launch more 'dummy' next-gen Starlink 2.0 satellites, which will have roughly four times the bandwidth of their earlier satellites. Currently, about 8500 Starlink satellites are operating in low-Earth orbit. As the new V2.0 satellites are four times heavier, SpaceX says that Starship will be the only vehicle capable of mass-launching them.

A previous Starship mission demonstrated something that's been likened to a Pez dispenser (a character toy that pops out Pez lollies, a sort of compressed sherbet). It ejected a series of dummy units the same size and weight as the new V2.0 satellites, like slices of toast popping out a toaster, as shown in time lapse at https://youtu.be/ZUKi2IVTwKc

According to SpaceX, certain thermal shield tiles have also been removed to test some vulnerable areas when Starship deorbits and splashes down in the Indian Ocean. While some may goggle at the sight of a rocket plunging into the sea, as though the mission has failed, Starship's eleventh flight is essentially another crash test in a long line of trials designed to capture more data.

The ability of the Super Heavy booster to land vertically and be captured by 'chopsticks' is an astonishing spectacle, which can be seen at https://youtu.be/RYUr-5PYA7s

Lunar power

Meanwhile, NASA's Artemis SLS (Space Launch System) will ultimately take people back to the moon; a crew of four is due to slingshot past the lunar body on the next Artemis mission.

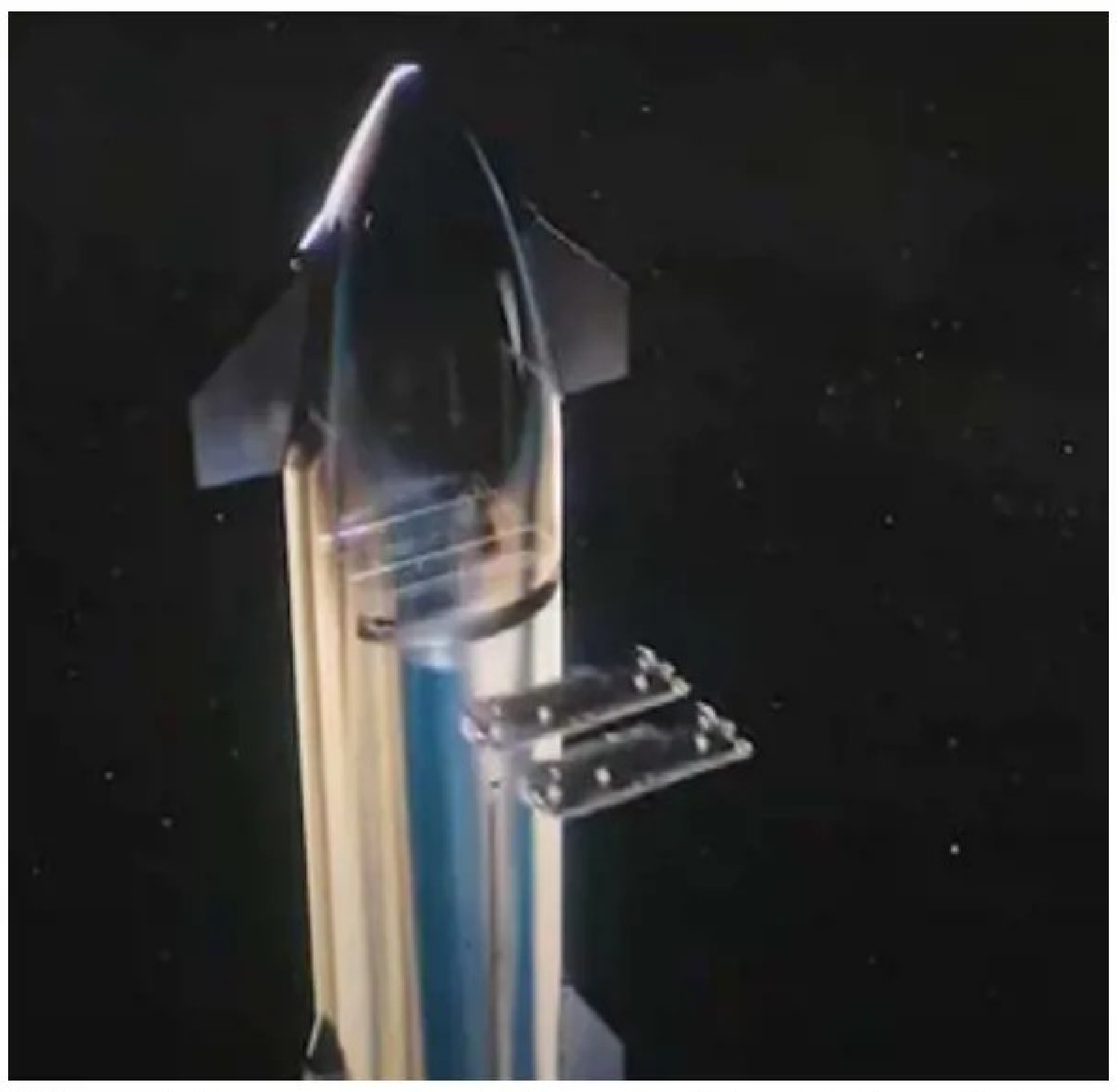
As Artemis aims to establish a permanent US presence on the moon, NASA recently published their *Directive on Fission Surface Power (FSP) Development*, which proposes accelerating the construction of nuclear fission micro-reactors both on the moon and, eventually, on Mars.

They cite the need to support a future 'lunar economy', generate high-power energy on the lunar and Martian surfaces, and strengthen America's national security in space.

The NASA Directive pulls no punches in expressing a newfound urgency for the project, warning that China and Russia may jointly build a nuclear plant on the moon by the mid 2030s, and then attempt to establish 'keep out' zones that block the US from doing the same (think of the hotly contested South China Sea).

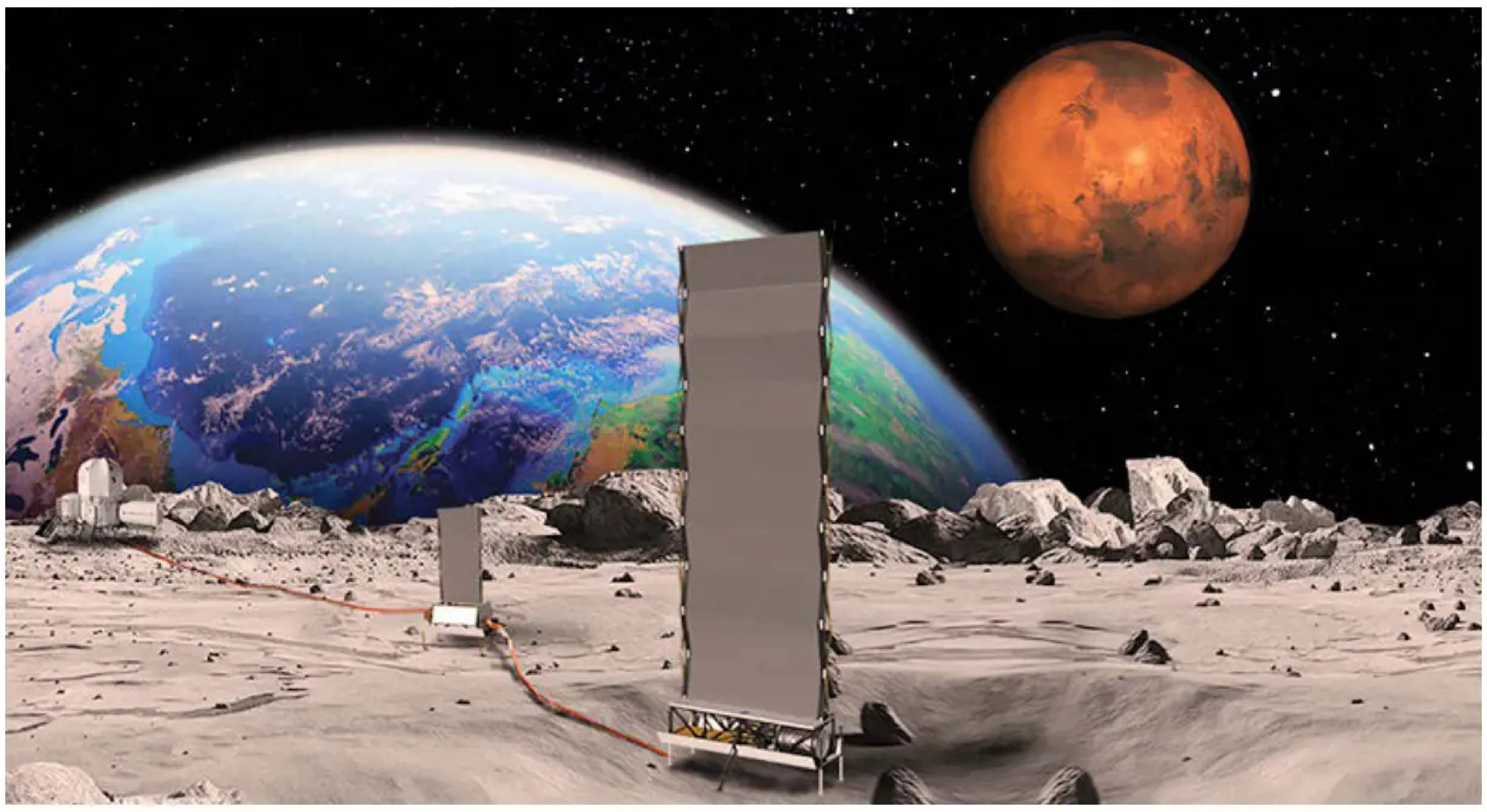
A new space race is in the making, as the US endeavours to be the first to build a nuclear power plant on the moon within the next five years.

Back on Earth, there are increasing concerns that China is striving to take the lead in the emerging nuclear fission sector. The country is reportedly harnessing laser inertial fusion (see https://w.wiki/FXhw) and existing technologies in an effort to leapfrog anything currently available in the West.



Starship launching Starlink V2.0 satellites into low-earth orbit, using a system dubbed the 'Pez dispenser'.

SpaceX's Starship is the largest rocket ever built. It is being 'crash tested' into the Indian Ocean after gathering more test flight data.



A concept image of NASA's fission surface power (FSP) Project, a nuclear power plant destined to be constructed on the moon. Source: NASA.

The classic two-pole SAE connector.

Chinese scientists recently claimed a world record for a fully superconducting magnet, which could ultimately be harnessed to generate extremely strong magnetic fields for applications in nuclear fission, for example. This is a sign of how urgently the country is researching all aspects of generating nuclear fusion energy, with a view to entirely dominating the market.

Charging up for winter

Back home in England, with wintertime gradually approaching ("winter drawers on", as my old aunt used to say), attention is drifting towards the matter of maintaining certain types of battery and battery-powered equipment through the cold season.

Most car owners know how a vehicle's lead-acid battery can be punished by the heavy use of heated screens, demisters and cabin heaters, with short journeys, heated steering wheels, seats and lighting all adding to the battery's load.

The judicious use of stop-start systems can also hammer even the best absorbed glass mat (AGM) car batteries (which is why I disable stop-start anyway).

Depending on your lifestyle, if car usage is only occasional or intermittent, then a maintenance trickle charge through winter will keep a battery primed and ready for service. This also helps avoid the risk of a battery self-discharging through neglect, and possibly becoming damaged beyond recovery.

The problem of maintaining batteries during winter can affect not only car owners but also those who own, say, a motor home, motorcycle or classic car. If you're going to leave a battery on charge long-term, you need to make sure the charger is supplying the correct float voltage, so it isn't over-charging or under-charging the battery.

Popular smart charger brands include Sweden's CTEK and the US manufacturer NOCO. During demanding conditions (cold winter weather, short journeys, heavy loads etc), the Author sometimes trickle-charges his car battery in situ in the garage, either using a special CTEK connector permanently fitted under the battery posts, or by hooking up crocodile clip leads instead.

The Lawnmower Man

Next in line for winter treatment was a small 12V sealed lead-acid (SLA) battery that's used to power a lawnmower starter motor. Although new battery chemistries come along all the time, there is plenty of life left in the rugged SLA, with millions of them in constant use today.

In my own application, the best known battery type is the Enersys Cyclon brand (see **https://pemag.au/link/ac8w**). In my experience, they have peerless reliability and performance - a Gates (now Enersys) Cyclon battery came fitted to the lawnmower when new and it endured many years of punishment.

For anyone interested, some notes and the application data for Cyclon batteries are downloadable in PDF format from **https://pemag.au/link/ac8x**

This type of compact SLA battery appears in garden machinery, tractors and other equipment, as the Cyclon's AGM design and chemistry offer a high energy density while withstanding vibration and rough treatment. These rugged batteries are easy to live with, and they also have a low self-discharge rate. Consequently, they're utilised in a very wide range of demanding applications.

If I need to replace the battery, I'll look for the six-cell module type (ie, 2V/cell), similar to the Cyclon, rather than imported block-shaped so-called 'leisure' batteries that perform badly. A compatible equivalent (though not a genuine Cyclon) that I'm currently using is available from Amazon at **https://pemag.au/link/ac8y**

When used in garden machinery or, say, motorcycle applications, one thing this style of battery has in common is the two-pole bullet connector, known as SAE (formerly the Society of Automotive Engineers).

The two-pole SAE is something of an enigma, as it seems to have evolved from automotive use over many decades, but there appears to be no specific design standard. These days, the SAE finds its way into low-voltage solar panel applications as well. The Wikipedia entry at **https://w.wiki/FXiG** sheds a little more light on the background.

Several electronic 'smart' battery chargers offer a variety of 12V connectors, including SAE types, which would cater both for my car and the lawnmower battery. The CTEK MXS 5.0, for example, is used by leading vehicle makers. CTEK sells a huge range of connectors, panel-mounting sockets and extension leads to go with their chargers.

At one time, I hard-wired a trailing CTEK socket to my car battery (a ring terminal goes under the battery post clamp nut), so I could simply lift the bonnet (hood) and hook up the charger when I wanted to top it up overnight.

What piqued my interest is that the SAE connector is of course keyed for polarity, which I found becomes very relevant when using a smart charger to charge a compatible battery. If you hook this style of battery directly to such a charger,

MXS 5.0
CTEK
12V/5A
MODE
RECOND

The CTEK MXS 5.0 is a popular 12V smart charger used by vehicle manufacturers, but note the polarity of the SAE connector.

the polarities are reversed, which technically shorts out the charger; its protection circuit is relied upon to detect the fault.

More recently, I tried a small mains-adaptor style 'smart charger' produced by Sealey Tools. Their Autocharge100HF is a 1A 6V/12V compact charger that comes with a few connectors, including the SAE type. It had the same battery polarity issue – plugging them together will short the charger.

As shown in my photo, what's needed is an SAE polarity changer, which is widely available from AliExpress etc. It seems this workaround is common knowledge in automotive circles, but nowhere did battery charger manufacturers point it out.

CTEK was initially unaware of this anomaly until I sent them photographs. I found Sealey's operating instructions were not the best, and Sealey did not respond to my query.

The lesson is clear: double-check the polarity is correct when connecting these types of products.

Looking for Live wires

Regular *Net Work* readers will recall that I picked out some digital testers and multimeters that highlighted the low prices of some test gear that's now available from overseas.

Last month, I suggested the Aneng M109 "Smart Voice Broadcast" meter, a pocket-sized DMM that announces the readings and range settings. This feature helps when you're busy prodding around a circuit board with your test probes. After waiting for a discount offer or two, the M109 came in at a miserly £4.

Two more low-cost digital multimeters from the same manufacturer recently arrived that will offer hobbyists a choice of features and convenience. They may also appeal to younger enthusiasts or anyone operating with a modest budget.

If a talking multimeter such as the M109 isn't for you, a basic DMM is offered in the Aneng M113. This compact auto-ranging 1999-count digital voltmeter is rated up to 600V AC/DC, and it measures resistance up to 10MΩ. With only three buttons, it's very easy to use and will suit anyone who doesn't need advanced features.

The meter includes a continuity buzzer, data hold and non-contact voltage testing. The NCV function worked well enough, but is probably best used as a guide only. Both red and black housings are available, and the price weighs in at less than £5 on AliExpress and Temu. Some keen reviewers say they buy several at a time so that a basic multimeter is never far away.

The Aneng 620A sports a large, easy-to-read LCD screen with digital and bargraph segments. As well as the usual ranges, it measures lower values of capacitance and frequency, and a thermocouple bead (supplied) can be used to measure temperatures up to a claimed 1000°C.

A polarity changer is needed to charge an SLA battery from a smart charger.

The instructions aren't as clear as they could be, I felt, and a mystery 'single watch pen wire test' is, in fact, a live voltage probe. Touching the positive probe tip to a live mains contact turns the LCD backlight red, and it flashes up a warning.

I liked the large screen display, which made it easier to view readings, but I found the NCV function somewhat hit-or-miss. A built-in torch (or 'flashlight') offers a little direct illumination.

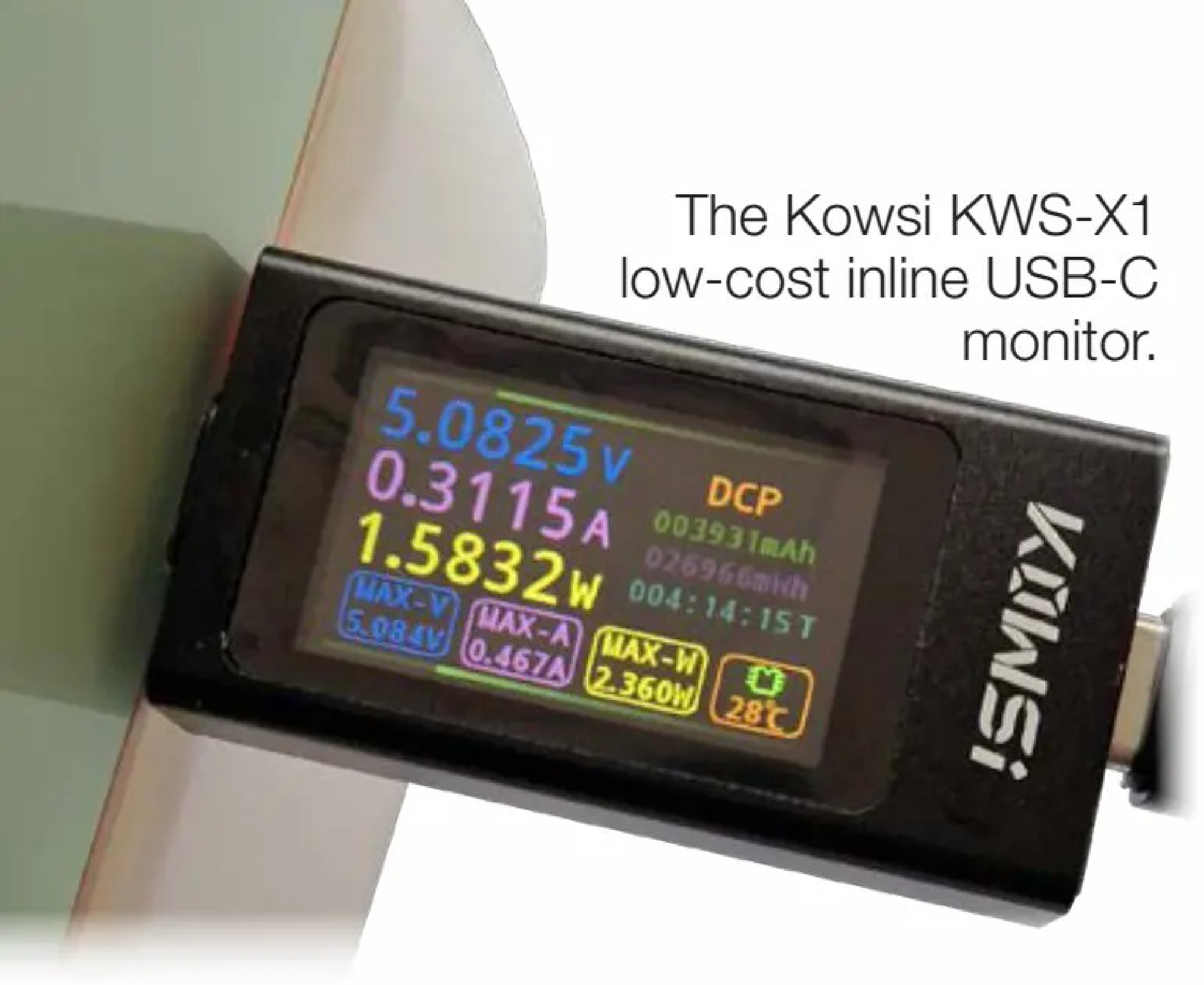
The socket legends are displayed on the LCD screen rather than the meter body itself, and a silicone rubber sleeve guards against knocks. You need to look closely to determine the button legends, though. The case has no tilt prop.

The LCD would not survive a direct hit on the workbench, but Aneng packs a lot in for such a low cost. I've seen it listed at up to £35, but my model arrived for under £6. Look out for extra kits bundling a full set of test leads, available from the usual sources.

Testing protocols

I've previously mentioned a range of USB adaptors sold under the Kowsi brand that contain a digital display to monitor current, voltage and more besides. I find them useful for checking progress when charging power banks, tablets and smartphones. When Power Delivery (PD) is used for faster charging, this is also denoted on the display.

The Kowsi KWS-X1 is a USB-C dongle that measures ripple voltage, current and PD/QC protocols, showing digital data or a 'scope-like curve on its colour IPS display. The sampling speed is selectable from thirty seconds to two hours.



The Kowsi KWS-X1 low-cost inline USB-C monitor.

It's feature-packed and rated at 4-30V/0-12A, but getting to know the device takes some effort.

I use one just to display a power bank charging curve at a glance over a few hours, but some enthusiasts are already hacking the firmware to fix minor bugs. The Kowsi KWS-X1 is about £12, but mine came from Temu for just £6. If you need to check out charging protocols, this low-cost USB-C inline tester might be worth considering.

Windows 10 updates – a respite!

Recently, while I was glued to the television, an advertisement appeared for the Currys electrical chain. They were very excited about selling so-called "AI PCs" and the ad played heavily on consumer fears that Windows 10 would suddenly become vulnerable, as if overnight, once Microsoft halted any further security updates for this decade-old operating system.

Net Work isn't a computer column, but many readers will be affected by the fact that the official cut-off date for Windows 10 security updates was October 14th, 2025. Back in September 2021, I wrote that Windows 11 was coming over the horizon but, because of stringent hardware requirements, not every PC would be able to upgrade to it.

The trusted platform module (TPM) is a device for storing security keys embedded on the PC motherboard. It is designed to nail down the PC's security. Without a TPM, Windows 11 could not be installed – that was Microsoft's ambition, anyway.

At the time, there was an outcry about the elite hardware specifications needed to run Windows 11. Predictably, the 'lack of security' of Windows 10 was used by the industry to fret users into buying new hardware, and the same thing is now happening all over again.

On many more recent machines, a TPM can be enabled in the computer's BIOS (basic input/output system) or by upgrading the BIOS version. A visit to the motherboard maker's website will confirm what's possible with your machine. On the author's home-brew Asus PC, simply by updating the BIOS, the system became Windows 11 compatible.

My computer colleague pointed me towards FlyBy11 (now called Flyoobe), a workaround that lets Windows 10 systems upgrade to Windows 11, whether or not they meet the hardware requirements. (Barry Fox also suggested it in the October issue.)

Flyoobe is donation-ware that installs a version of Windows Server to set up a 'normal' installation of Windows 11. It's probably a job best left to confident PC users with time on their hands; it can be downloaded from **https://github.com/builtbybel/Flyoobe**

Before toying with a Windows 11 upgrade, I discovered that a workaround had sneaked onto my system following a Windows update. Consumers using a Windows 10 PC can now enjoy one more year of critical security updates until 2026, which may be free of charge or cost $30.

You can also 'pay' for the extension by cashing in 1,000 Microsoft Reward Points (click the toolbar Search icon, and your Points balance shows along the top). Those who live in the EU don't even need to do that; Microsoft has agreed to let them have the year of updates for free.

Where a PC is eligible for Windows Extended Security Updates (ESU), an "Enrol Now" link will appear in the Windows Update section of Settings. A few basic conditions have to be met, but it's likely that most consumer systems that are up to date will automatically be eligible for ESU.

I followed the on-screen instructions to 'enroll' my PC to receive another year of security updates.

If I decide to upgrade to Windows 11, I shall use Macrium Reflect to create a disk image first on a spare disk, so that I can revert to the old OS if necessary. Otherwise, I'll install a clean OS upgrade onto a new SSD. You can read all about Windows 10 Extended Security Updates at **https://www.microsoft.com/en-GB/windows/extended-security-updates**

SEALEY
COMPACT AUTO SMART CHARGER
MODE
12V
6V

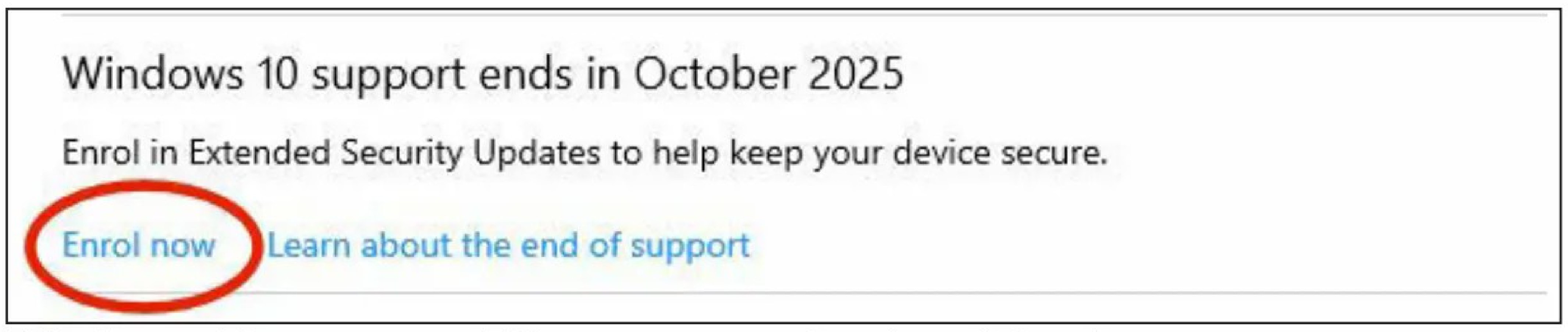
The Sealey Autocharge 100HF compact smart charger.

TASHIMA
ANENG
M113
DIGITAL MULTIMETER
8.2

The Aneng M113 pocket-size DMM.

ANENG
620A
Smart Multimeter
AUTO
Auto
20 °C
T-RMS
6000Counts
10A
COM
VHz
10A/10SEC MAX
DC/AC 600V CAT III

The Aneng 620A.


Windows 10 support ends in October 2025

Enrol in Extended Security Updates to help keep your device secure.

Windows 10 consumer PCs can now extend updates for one more year.

The future is… AI?

Given the way things are going in the personal tech world, to get the best out of Windows 11, a modern "AI" PC warrants serious consideration. There's no avoiding the fact that AI is increasingly having an impact on everyday workflows and PC usage. The days of googling for information are fast receding as users can now interact with AI in a human-like way.

Online, Amazon's Rufus AI chatbot produces eerily human responses when quizzed about a particular product. Increasingly, I have to stop and remind myself that I'm talking to ChatGPT, which costs anything from free to £200 a month – an app is available from the Microsoft Store.

Microsoft's Copilot has also found its way onto the Windows desktop and, when the microphone's 'hot', it will soon start burbling away.

I must admit that Microsoft Copilot has turned up trumps on more than one occasion. For example, using some legacy email software, I wanted to configure some arcane port settings for use with a new service provider. I was stumped for the answer, but Copilot immediately pointed me to a file deep in the software along with instructions to drag and drop a certain file into the main program folder.

Astonishingly, this worked immediately, and the port settings became accessible. Copilot also reminded me of some Excel spreadsheet shortcuts.

I'm also expecting to move towards using voice recognition for drafting written copy. Windows 10 users will find options in the Ease of Access area of Settings, but my experience with a headset was very poor. Windows 11 'AI' PCs have a neural processor unit (NPU) optimised for such tasks.

Suffice it to say that my computing counterpart, who is seldom easily impressed, had nothing but praise for a client's Windows AI laptop that was using voice recognition. He was actually "blown away", so perhaps that's something to go on my Christmas wish list. One thing's for sure, AI for desktop and web users is here to stay, and it will only become more prevalent – and intrusive – over time.

Bye Bye, Mr. Babani

I'm very grateful to reader Andrew Gale, who dropped me a line with some news:

*I've always enjoyed your pieces in PE and posted on your website (**alanwinstanley.com**), especially your historical/nostalgic writings about hobby electronics.*

I couldn't help but notice that Babani Publishing has recently wound-up. I always loved their books, which were a big part of my introduction to electronics (along with Everyday Electronics and books by Forrest M. Mims III). I was surprised that their closure seems to have passed without comment, and wondered if you knew anything about it.

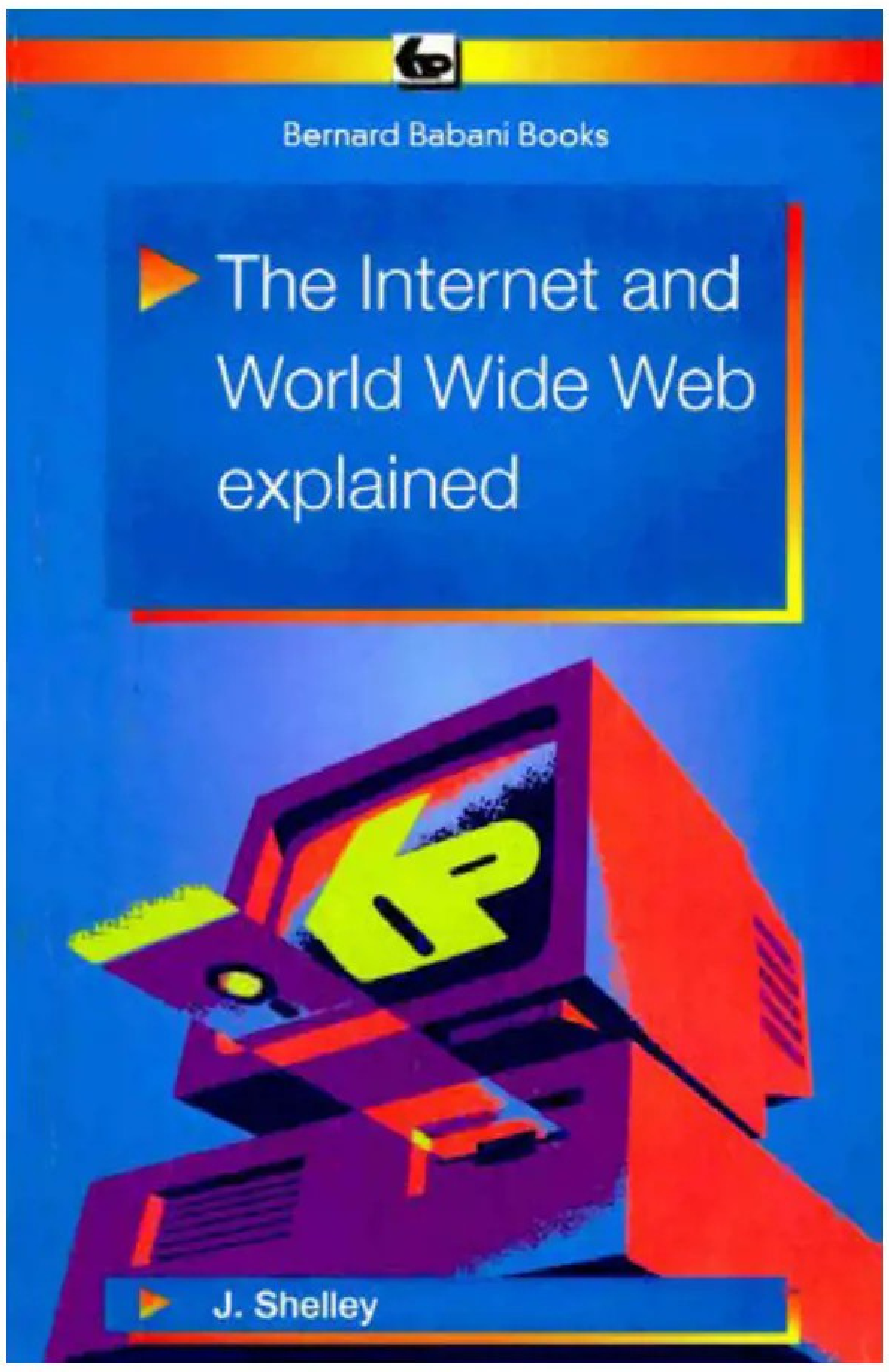
I guess Michael Babani must have reached retirement. Thanks for maintaining a fascinating website and, if you do know anything about Babani, I'd love to hear!

British electronics constructors of a certain age (mine) will fondly remember the pocket-money paperback books published by Bernard Babani over many decades.

London-based Babani Electronic Books found a place on every hobbyist's bookshelf at one time or another, in an era when hobbyists relied completely on the printed word as a source of essential technical data, project circuit ideas and inspiration in general.

Eventually, hundreds of modestly priced Babani books would embrace technical topics, software, microprocessors and the web, including the 1996 title *The Internet and World Wide Web explained*, published in the same year that *Net Work* was founded.

Many regular and well-known PE/EE magazine writers contributed to the catalogue of Babani titles, including Robert Penfold, the late Frank (F.G.) Rayer (who was also a science fiction writer), R. N. Soar, E. A. Parr, O.N. Bishop and D. W. Easterling amongst others; names that will be familiar to old hands.



This 1996 book was typical of the ubiquitous handbooks published by Bernard Babani.

The Wiki page at **https://pemag.au/link/ac8z** has a lot of interesting background about its founder, Bernard Babani. I was not aware that a teenage Clive Sinclair (1940-2021) ran the firm's publishing side for several years.

The publisher's cryptic address of "The Grampians" in Shepherd's Bush Road, London, conjured up images of a mysterious place that seemed a world away for young Babani readers like myself. Thanks to Google Street View, it can now be revealed in all its glory: it's a ground-floor business property under an apartment block (see **https://pemag.au/link/ac90**).

Reader Andrew visited The Grampians in London and emailed me some up-to-date photos! I wonder what Bernard Babani would have thought about that idea.

Enquiries show that Bernard Babani Publishing was wound down earlier this year after, and it's safe to assume, its dedicated director, Mr Michael Babani, the son of Bernard, finally retired. Sadly, this means that the book catalogue is likely to go out of print as a matter of course, but used copies will appear on the likes of **abebooks.co.uk** and Amazon.

So, on behalf of readers and hobbyists everywhere, I'd like to wish Mr Babani a very happy retirement, and a sincere thank you from every one of us, for being a staple part of our adventures in electronics that inspired us for more than half a century.

That's all for now – join me next time for more *Net Work*. You can contact me at **alan@epemag.net** PE

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BATTERY-POWERED MODEL TRAIN

BY LES KERR

This modification eliminates the need to keep model railway tracks clean. If you let them oxidise, power won't get to the trains, causing all sorts of problems. By making the train battery powered, it no longer needs to draw power from the tracks, making it much more reliable!

My grandson was visiting and he was at me all the time to let him drive trains on my OO gauge railway. As it hadn't been used for quite a time, there was quite a build-up of debris on the track and the engine pickups that resulted in the first train running erratically. After laboriously cleaning the track, the trains ran smoothly.

Most of my newer power tools are battery powered, so I wondered if I could power the train from onboard rechargeable AAA cells. These could be mounted in the carriage behind the engine, and the speed and direction could be controlled by a simple 433.9MHz link. I calculated that four fully-charged 900mAh NiMH cells in series could run the train for more than five hours on a charge.

With most model train layouts, the 433.9MHz transmitter will only be a few metres from the train at any time, so there is little chance of interference. To ensure the train doesn't go haywire, check bytes are sent so that the Receiver can verify the speed and direction data it is getting are correct. This virtually eliminates the possibility that erroneous signals will result in incorrect operation.

The NiMH battery voltage of around 4.8V is too low to run the motor, so I selected a small step-up converter module that produces 15V DC from the battery voltage to power the motor. It operates at about 1MHz with an efficiency approaching 90%. Another small boost converter generates a steady 5V rail to run the control circuitry.

To drive the engine, I looked at all the standard H-bridges on the market and selected the DRV8871 IC that is mounted on a 24.5 × 20.5mm PCB. It runs from 5-37V at up to 2A, driving a single motor bidirectionally. Over-temperature and over-current protection is built in. It is a bit overkill for a 12V train that takes about 250mA maximum, but it could be used with higher power engines too.

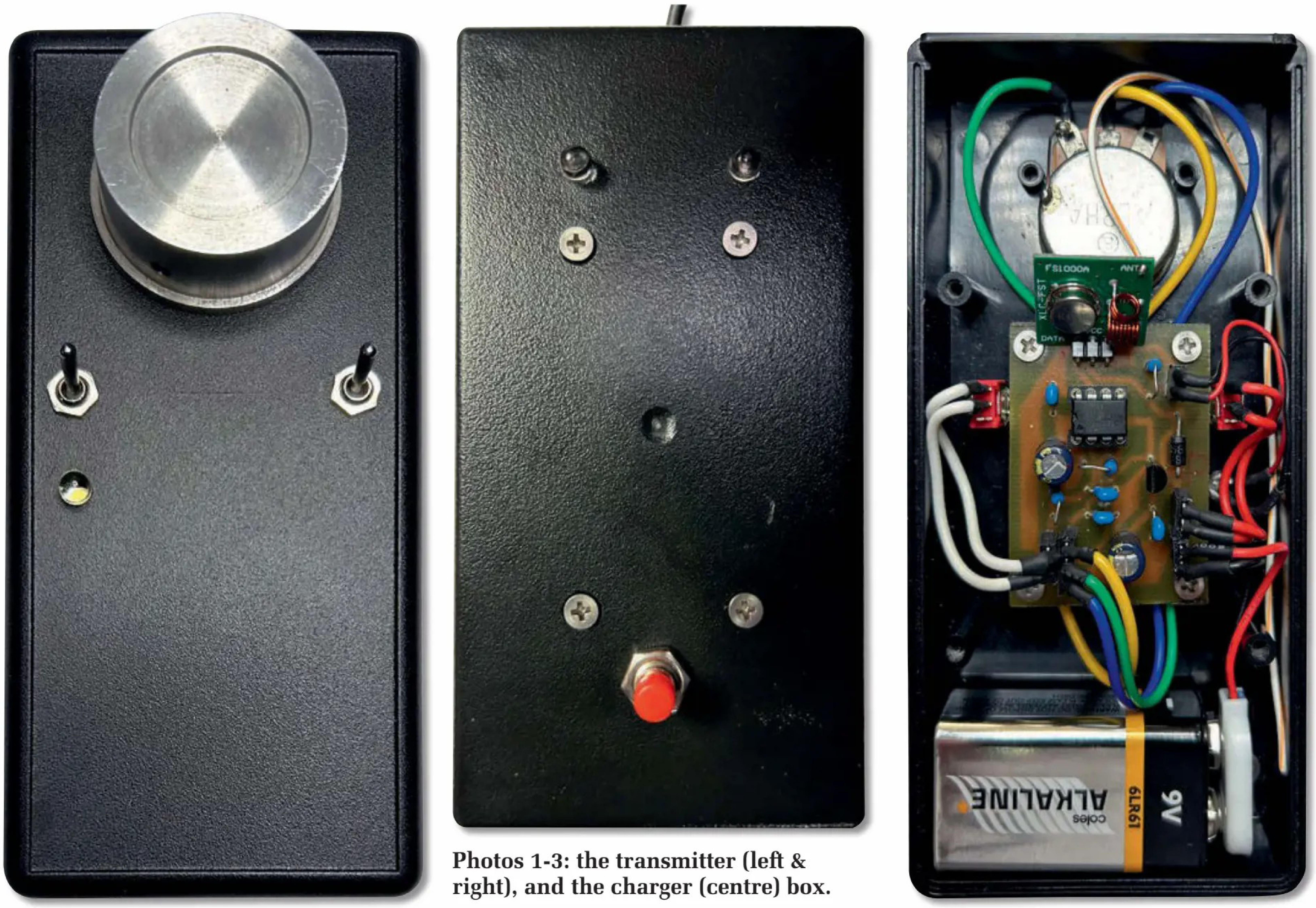
The motor speed and direction are controlled by a microcontroller on the same PCB as the motor driver that mounts in the carriage, behind the engine. This PCB also has a 433.9MHz receiver to allow remote control. To cater for various size carriages, I designed two Receiver PCBs, a small one using SMD components (carriage length 85mm) and a larger one with through-hole (TH) components.

The handheld controller (Photo 1) has a potentiometer that controls the speed of the train and a toggle switch to select forward or reverse. The Transmitter has a PIC12F617 microcontroller that monitors those controls and sends signals via a 433.9MHz transmitter within the handheld controller.

A 3mm red LED on the carriage lights when the battery needs charging.

The fourth PCB I designed is a trickle Charger (Photo 2) that connects to a socket on the battery carriage using a 2.5mm jack plug.

This system of three modules – Transmitter, Receiver and Charger – provides everything you need to run a


ALKALINE
coles
6LR61
9V
FS1000A
ANT

Photos 1-3: the transmitter (left & right), and the charger (centre) box.

model locomotive without requiring an electrical connection (for either power or communications) through the track. You can see a video of it in operation at **siliconchip.au/Videos/Battery+model+train**

Transmitter circuit details

The Transmitter circuit is shown in Fig.1. It is powered by a 9V battery via on/off toggle switch S1 and a 1N5819 schottky diode. The diode prevents accidental battery polarity reversals from destroying the circuit. A schottky diode is used as its forward voltage drop is a lot less than a standard silicon diode, so the battery lasts longer.

A 78L05 regulator provides +5V for the microcontroller. The 100µF capacitors connected to its input and output reduce any ripple to a negligible level, while the 100nF ceramic or MKT capacitors reduce any high-frequency noise that may be present.

So that potentiometer VR1 varies the train speed, microcontroller IC1 measures the voltage at its wiper using its internal analog-to-digital converter (ADC) via analog input AN3. It converts the 0-5V on its wiper to an 8-bit number between 0 and 255. That value is sent out as pulses via digital output GP0 (pin 7), to the transmitter module, to be picked up by the Receiver on the train.

Digital input GP5 (pin 2) is pulled

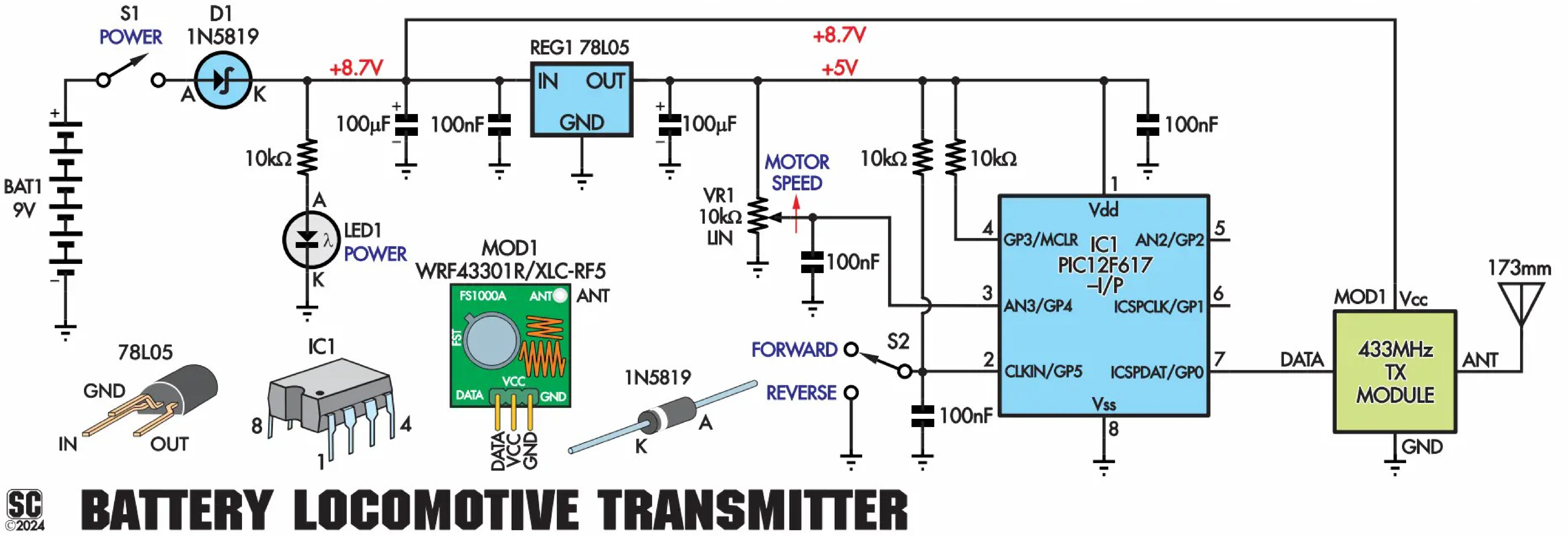

S1
POWER
D1
1N5819
A
K
+8.7V
REG1 78L05
IN
OUT
GND
+8.7V
+5V
BAT1
9V
100µF
100nF
100µF
100nF
10kΩ
LED1
POWER
MOTOR
SPEED
VR1
10kΩ
LIN
10kΩ
10kΩ
100nF
Vdd
GP3/MCLR
IC1
PIC12F617
–I/P
AN2/GP2
AN3/GP4
ICSPCLK/GP1
CLKIN/GP5
ICSPDAT/GP0
Vss
FORWARD
REVERSE
S2
100nF
MOD1
WRF43301R/XLC-RF5
FS1000A
ANT
DATA
VCC
GND
1N5819
78L05
GND
IN
OUT
IC1
8
4
1
MOD1
Vcc
433MHz
TX
MODULE
DATA
ANT
GND
173mm
SC ©2024
BATTERY LOCOMOTIVE TRANSMITTER

Fig.1: the Transmitter circuit. It runs from a 9V battery; microcontroller IC1 and transmitter MOD1 convert the position of speed potentiometer VR1 and forward/reverse switch S2 into a 433.9MHz-modulated ASK serial data stream for the Receiver.

high by the 10kΩ resistor when S2 is in the forward direction or low, to ground, by S2 when it is in the reverse direction. The microcontroller senses this level using its GP5 digital input and sends different numbers via the 433.9MHz transmitter depending on the switch state. The 100nF ceramic capacitors at those two inputs prevent noise from affecting the readings taken.

The signal sent to the transmitter module via the GP0 output is serial data at 1200 baud that contains the speed and direction variables, along with preamble and check bytes. This 433.9MHz module transmits this using amplitude-shift keying (ASK) via a quarter-wavelength (173mm long) wire antenna.

Receiver circuit

The Receiver circuit is shown in Fig.2. Signals from the Transmitter are received by the 433.9MHz receiver module, and the demodulated serial data is applied to the RC2 digital input (pin 8) of the PIC16F1455 microcontroller (IC2).

The 8-bit train speed data and the direction data are extracted and stored in memory, then used to generate the pulse-width modulated speed signal and the direction signal.

Two logic inputs, IN1 and IN2, control the H-bridge driver (IC3). To turn the motor in one direction, we apply a pulse-width modulated (PWM) signal to vary the speed to IN1 while holding IN2 high. If the train is to run in reverse, the PWM signal is applied to instead IN2 while IN1 is held high. To stop the train, both input are kept at the same level (both low or both high).

Fig.3: pulse-width modulation (PWM) involves setting the output high at a fixed interval, then leaving it high for a period ranging up to that interval. The result is a varying average voltage, even though the output only switches between two levels.

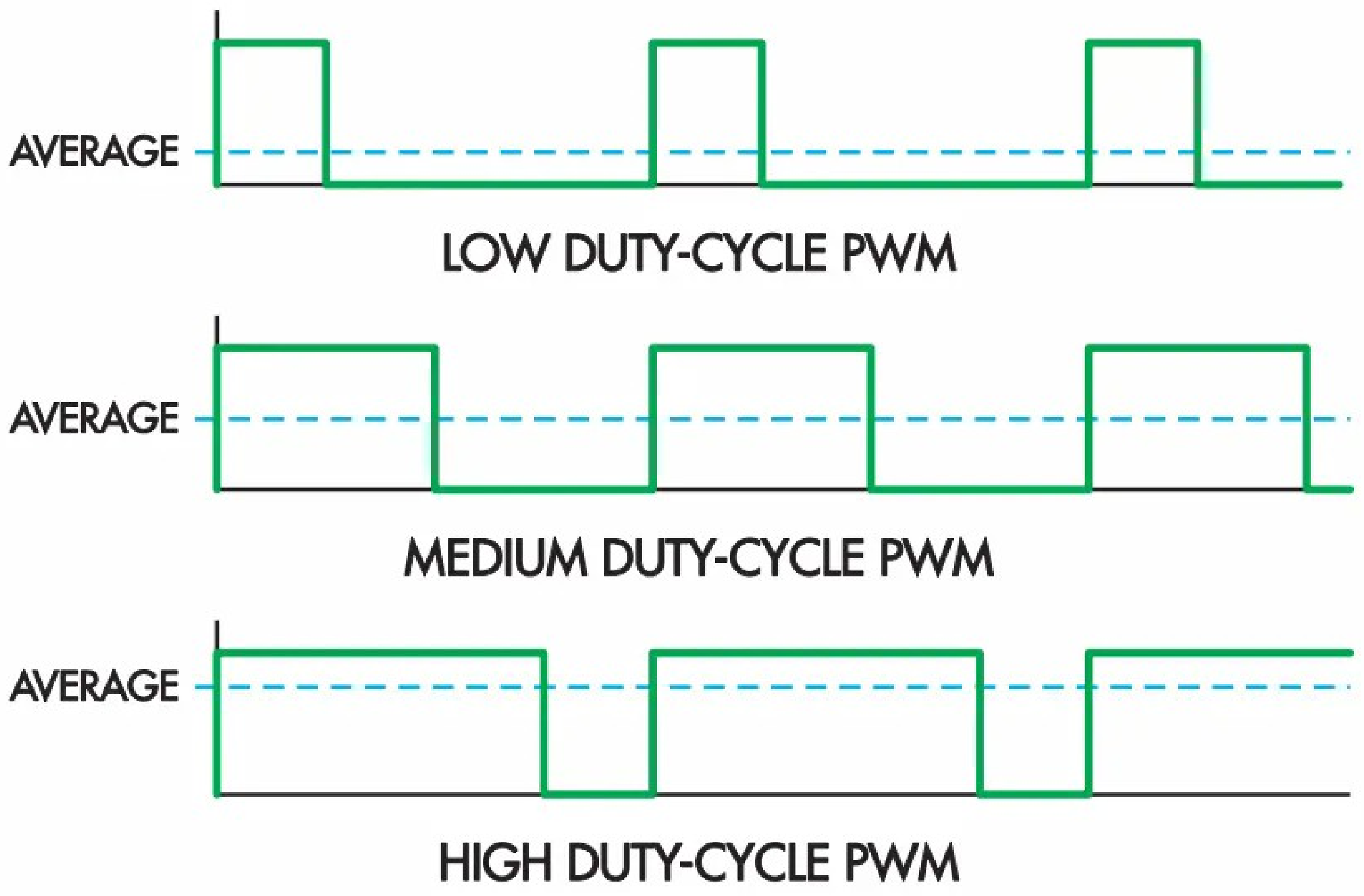


Fig.3 shows the signals for driving the motor at various speeds.

The battery supply voltage is halved by the two 10kΩ resistors and the resultant ~2.4V is monitored by analog input RA4 (pin 3) of IC2 using its internal ADC. If the voltage at that pin falls below 2V (ie, the battery is below 4V), digital output RC4 (pin 6) is taken low, switching on red LED2 to alert you that the battery needs charging.

The micro also provides signals to drive the DRV8871 H-bridge IC. To turn the motor in one direction, the PWM signal is applied to digital output RC3 (pin 7), while RC5 is taken high (+5V). To reverse the motor direction, the PWM signal is applied to RC5 and RC3 is taken high. The higher the speed value, the faster the motor turns.

When the speed control is near its minimum position, both RC5 and RC3 are taken low (to 0V), causing the PWM module to go into sleep mode, reducing the current drawn from the battery. The +5V supply for the receiver and micro is provided by the S7V7F5 high-frequency voltage up/down converter (MOD4) that takes the 4-6V battery voltage and provides a regulated +5V output.

If the battery has been recently charged (it could be as high as about 6V), MOD4 steps down the voltage to +5V; if it is discharged below 5V,

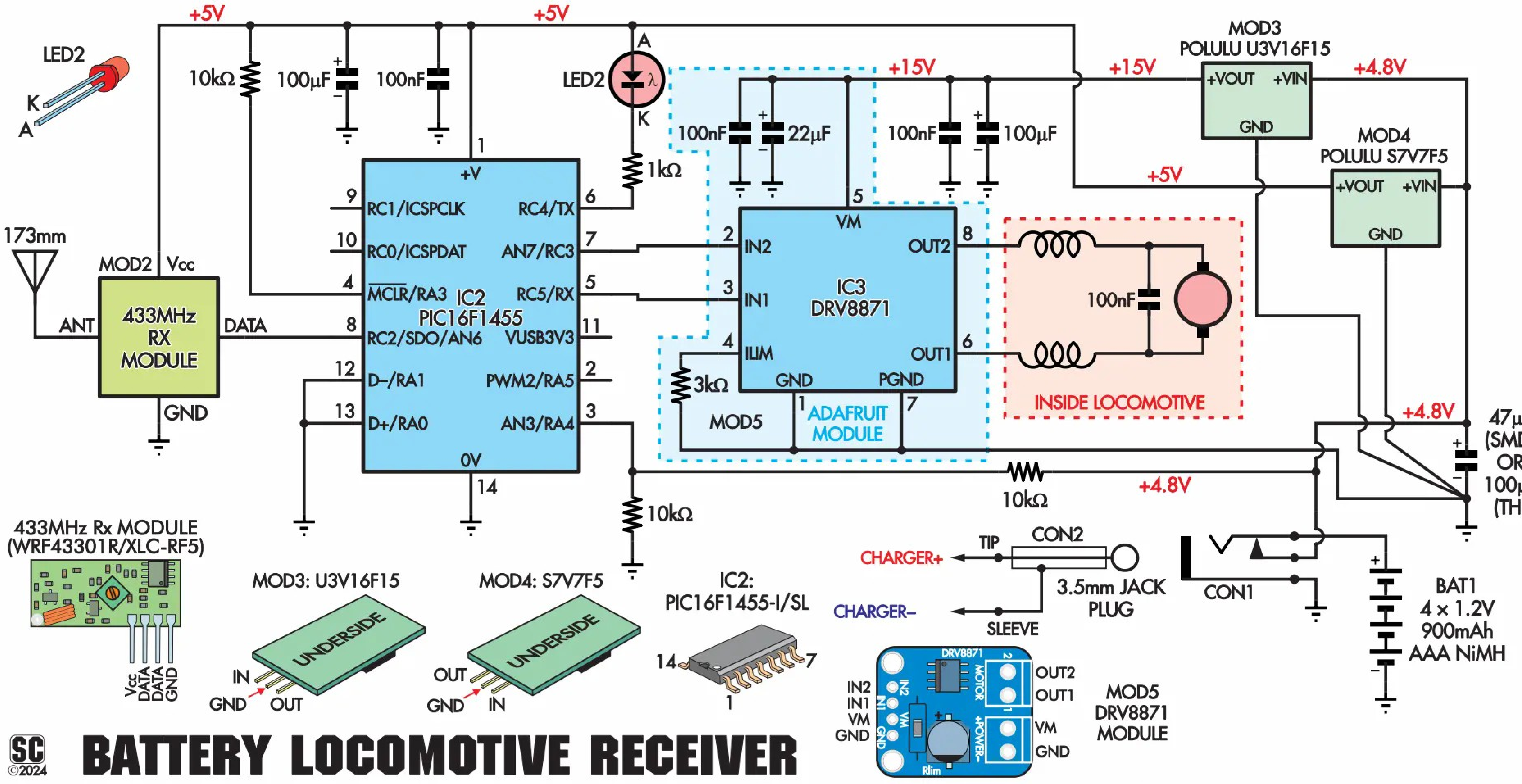

+5V
LED2
K
A
10kΩ
100µF
100nF
+5V
A
LED2
K
1kΩ
1
+V
9
RC1/ICSPCLK
10
RC0/ICSPDAT
4
MCLR/RA3
8
RC2/SDO/AN6
12
D–/RA1
13
D+/RA0
RC4/TX
6
AN7/RC3
7
RC5/RX
5
VUSB3V3
11
PWM2/RA5
2
AN3/RA4
3
IC2
PIC16F1455
0V
14
173mm
MOD2
Vcc
ANT
433MHz RX MODULE
DATA
GND
+15V
100nF
22µF
100nF
100µF
5
VM
2
IN2
3
IN1
4
ILIM
GND
PGND
OUT2
8
OUT1
6
1
7
IC3
DRV8871
3kΩ
MOD5
ADAFRUIT MODULE
100nF
INSIDE LOCOMOTIVE
+15V
MOD3
POLULU U3V16F15
+VOUT
+VIN
GND
+4.8V
MOD4
POLULU S7V7F5
+5V
+VOUT
+VIN
GND
+4.8V
47µF (SMD) OR 100µF (TH)
10kΩ
10kΩ
+4.8V
CON2
TIP
CHARGER+
CHARGER–
SLEEVE
3.5mm JACK PLUG
CON1
BAT1
4 × 1.2V
900mAh
AAA NiMH
433MHz Rx MODULE (WRF43301R/XLC-RF5)
Vcc
DATA
DATA
GND
MOD3: U3V16F15
UNDERSIDE
IN
GND
OUT
MOD4: S7V7F5
UNDERSIDE
OUT
GND
IN
IC2:
PIC16F1455-I/SL
14
7
1
IN2
IN1
VM
GND
OUT2
OUT1
VM
GND
MOD5
DRV8871
MODULE
SC
©2024
BATTERY LOCOMOTIVE RECEIVER

Fig.2: MOD2 picks up the data from the Transmitter and feeds it to microcontroller IC2, which decodes it and produces PWM waveforms for H-bridge motor driver IC3 on MOD5. MOD3 boosts the battery voltage to 15V to run the motor. IC2 also monitors the battery voltage and lights LED2 if it is low.

it steps it up. The 100µF electrolytic capacitor and 100nF ceramic capacitor reduce any noise or ripple on the supply. Similarly, the U3V16F15 (MOD3) provides the +15V DC supply for the motor.

We use 15V instead of 12V to overcome any voltage drop in the tiny cables connecting the carriage to the train motor. Polulu recommend in their data sheet that you add a 47µF capacitor across the battery input when using these inverters, which I have done. Both these modules are available online for around £7 each (eg, see https://uk.robotshop.com/).

There is a 2.5mm switched jack socket (CON1) so the battery can be charged. It also allows the battery power to the Receiver to be switched off simply by inserting a jack plug. With the jack plug in the socket, the battery is connected to the Charger and disconnected from the Receiver as its positive side is disconnected.

Charger circuit

Looking at Fig.4, the battery is trickle charged at C/10 (90mA) for 16 hours unless its output voltage exceeds 6V, indicating the battery is fully charged. In that case, the charge current is switched off. When the power pack is switched on, 9V is applied to the 78L05 voltage regulator (REG2), which reduces the voltage to +5V to power the PIC12F617 microcontroller, IC4.

The two 100µF capacitors smooth out any residual ripple, while the two 100nF capacitors provide high-frequency bypassing.

On powering up, digital output GP4 (pin 3) of IC4 pulses the green LED at 200ms intervals, indicating it is in standby mode. Pressing the Start button (S3) pulls the GP2 digital input low (pin 5), causing an interrupt routine to be triggered that takes the Charger out of standby mode and puts it into charge mode. The 100nF capacitor eliminates any contact bounce from the pushbutton.

This results in the green LED switching off and the red Charge LED flashing at 500ms intervals. Mosfet Q1 (IRL540N) is switched on by digital output GP5 going high, and the 16-hour countdown timer starts.

When on, the drain of the Mosfet goes low, connecting the 90mA constant current source to the battery. The current source comprises the BD136 transistor (Q2), an LM285 2.5V reference diode and a 220Ω resistor in parallel with a 22Ω resistor.

It works by holding the PNP base 2.5V below the +9V supply. This sets the emitter at 1.8V (2.5V – 0.7V), which matches the voltage across the parallel resistors. They have a resistance of 20Ω (220Ω || 22Ω). With 1.8V across 20Ω, Ohm's law (I = V ÷ R) tells us the current must be 90mA (1.8V ÷ 20Ω).

The battery voltage is halved by the two 10kΩ resistors and applied to analog input GP0 (pin 7) of IC4. Once per second, it measures the voltage and if it is above 3V (battery fully charged), charging stops and the Charger goes back into standby mode, shown by the green LED flashing. If the battery voltage doesn't exceed 6V, the charging stops after 16 hours.

The 1N4004 diode (D2) prevents the battery from discharging if it is left connected when the charger is not powered.

The 1N4148 diode (D3) prevents the ADC input from rising above 5.6V, although that is unlikely because the battery would have to be charged to over 11V. Still, it's possible CON2 could accidentally be connected to a voltage source, so it's better to be safe.

Sourcing parts

The receiver and transmitter modules are available from several suppliers under different part numbers. I

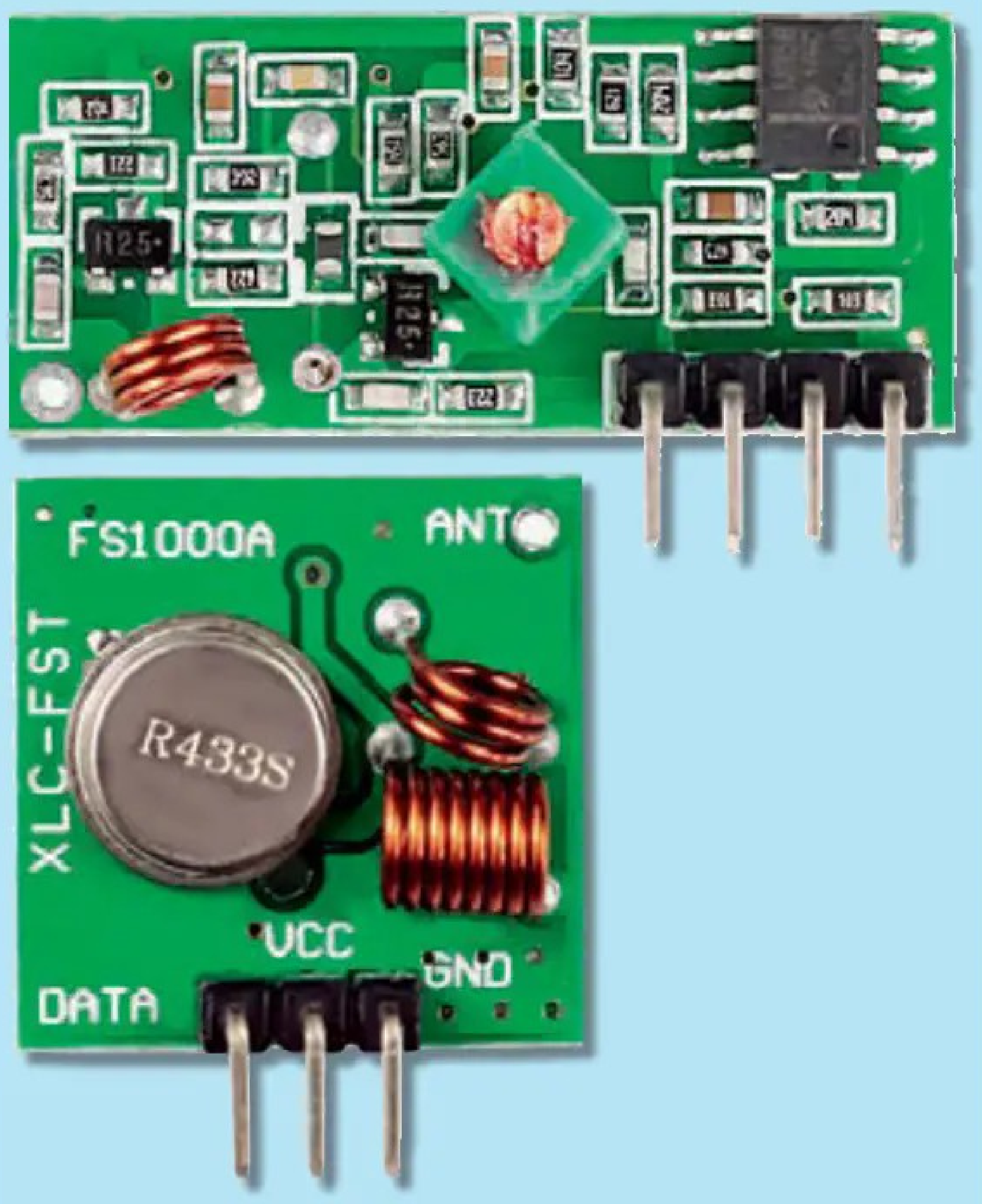
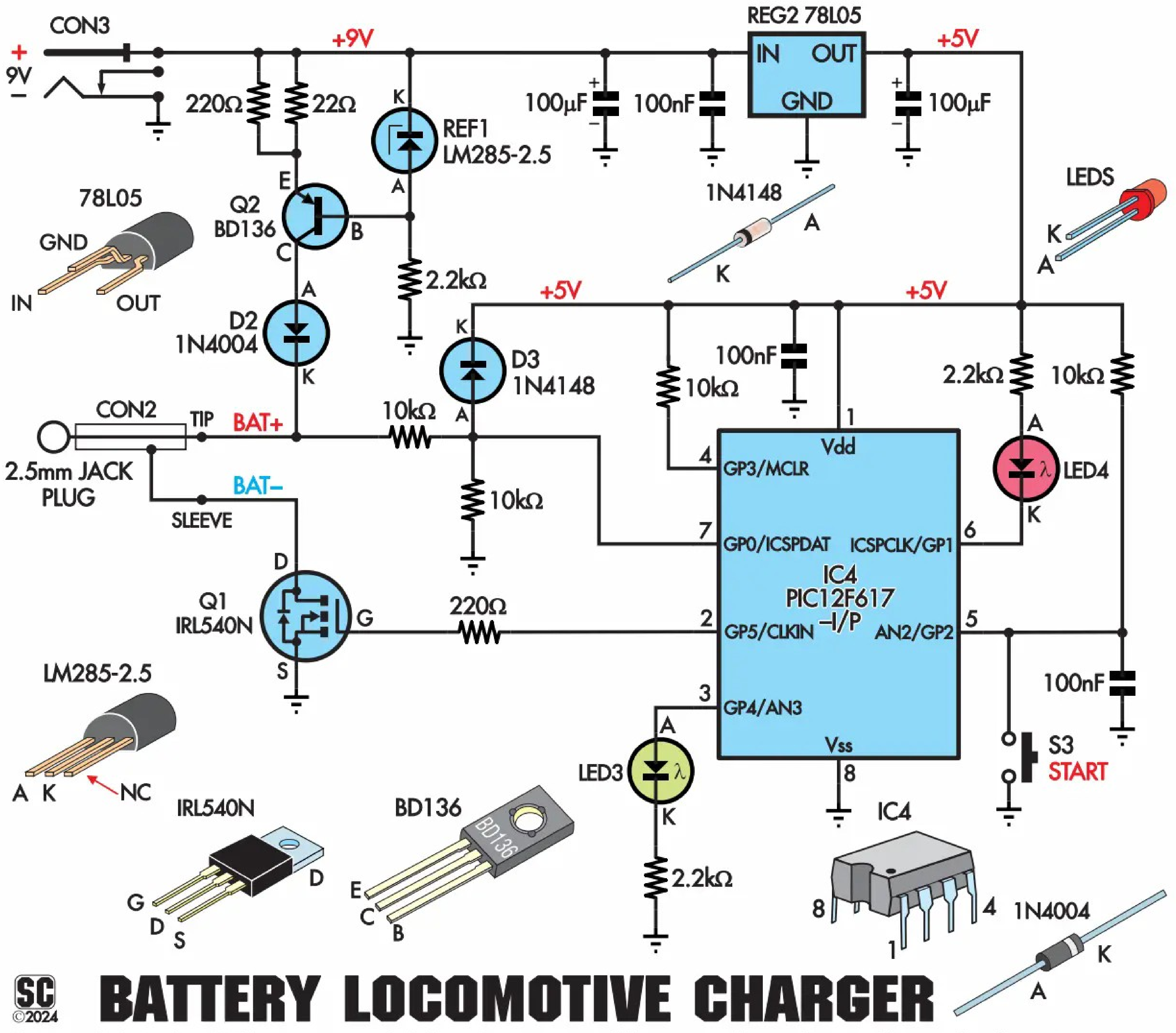

FS1000A
ANT
XLC-FST
R433S
VCC
DATA
GND

Photo 4: the 433.9MHz receiver (above) and transmitter (below) modules. They are sold under various model numbers, but this particular set is very common to find online. As long as yours look like these, and don't have low-quality soldering, they should work (avoid the cheapest ones!).


CON3
+
9V
–
+9V
220Ω
22Ω
K
REF1
LM285-2.5
A
100µF
100nF
REG2 78L05
IN
OUT
GND
+5V
100µF
LEDS
K
A
78L05
GND
IN
OUT
Q2
BD136
E
B
C
1N4148
A
K
2.2kΩ
+5V
+5V
A
D2
1N4004
K
K
D3
1N4148
A
100nF
10kΩ
2.2kΩ
10kΩ
CON2
TIP
BAT+
10kΩ
1
Vdd
4
GP3/MCLR
A
LED4
K
2.5mm JACK
PLUG
BAT–
SLEEVE
10kΩ
7
GP0/ICSPDAT
ICSPCLK/GP1
6
D
Q1
IRL540N
G
S
220Ω
2
GP5/CLKIN
IC4
PIC12F617
–I/P
AN2/GP2
5
LM285-2.5
3
GP4/AN3
100nF
A K
NC
LED3
A
K
Vss
8
S3
START
IRL540N
BD136
IC4
G
D
S
D
E
C
B
2.2kΩ
8
4
1
1N4004
K
A
SC
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BATTERY LOCOMOTIVE CHARGER

Fig.4: this NiMH battery trickle charger will stop charging when the battery voltage reaches 6V (1.2V per cell) or after 16 hours of charging. Q2, REF1 and the surrounding components form a 90mA constant current source while Mosfet Q1 controls whether charging is active.

Programming a microcontroller in-circuit

To program the micro with it in the circuit, you will need to solder wires to the +5V and 0V rails as well as pin 4 ($\overline{\text{MCLR}}$), and the pads on the ICSPCLK and ICSPDAT pins. Those are pins 9 & 10 respectively for the PIC16F1455, or pins 6 & 7 respectively for the PIC12F617. Connect those wires to your programmer, referring to its manual to see which wire goes to which pin. For the PICkit 3, the pins are (starting from pin 1) $\overline{\text{MCLR}}$, V_{CC}, GND, ICSPDAT and ICSPCLK.

You can download and install the free MPLAB IPE software from the Microchip website and then use the included MPLAB IPE software to open the appropriate HEX file (which you can download from siliconchip.au/Shop/6/508) and flash it onto the target chip via your programming hardware.

have given a couple of examples in the parts list, but there are many others. Sometimes the part number is for a transmitter/receiver pair and the individual parts don't have individual codes (or they are not specified).

The main thing is to check that what you are buying looks like the modules shown in Photo 4. If you type "433MHz modules" in a search engine, you will find plenty of suppliers of modules that look identical or nearly so. Be careful, though, as I found that one of the very cheapest suppliers' modules were poorly soldered and were unusable.

Construction

Let's start by building the Transmitter. It is assembled on a single or double-sided PCB coded 09110241 that measures 49 × 36mm. During assembly, refer to the PCB overlay diagram, Fig.5.

Fig.5 shows the off-board components wired directly to the PCB. You can do it that way, but it's easier to instead solder pin headers in those positions and then cut pairs of female-female DuPont jumper wires in half. That way, you can plug them into the headers and solder the bare ends to the other components.

You can see from the photos that I soldered wires to header sockets instead of using DuPont wires; either approach can work, but it's easier and slightly neater to cut the jumper wires in half. You can often get them joined together in a ribbon, making it easy to split off pairs or sets so they stay together (like a figure-8 cable).

Start the PCB assembly by fitting the headers, 8-pin IC socket and the capacitors. The IC socket makes it easier to remove the microcontroller and reprogram it later if necessary. Take care to orientate the socket and electrolytic capacitors correctly. For the electros, the longer positive lead goes into the pad nearest the + symbol, with the stripe on the can indicating the negative end opposite that.

Now add the resistors, which are mounted vertically, then the 78L05 voltage regulator, 1N5819 diode (with its cathode stripe facing as shown) and the 433.9MHz transmitter module. As the clearance inside the Hammond box is less than the height of the 433.9MHz module, the module should be mounted 20° from vertical towards the edge of the board (it's shown as if it's laid flat in Fig.5 for clarity).

Make sure all the semiconductors and the transmitter are correctly orientated. Don't fit the PIC12F617 microcontroller yet.

If you have purchased it from the Silicon Chip Online Shop, it will already have the firmware loaded. If you wish to program it yourself, you can download the firmware from: siliconchip.au/Shop/6/508

To load the firmware onto the chip, you will need a suitable programmer

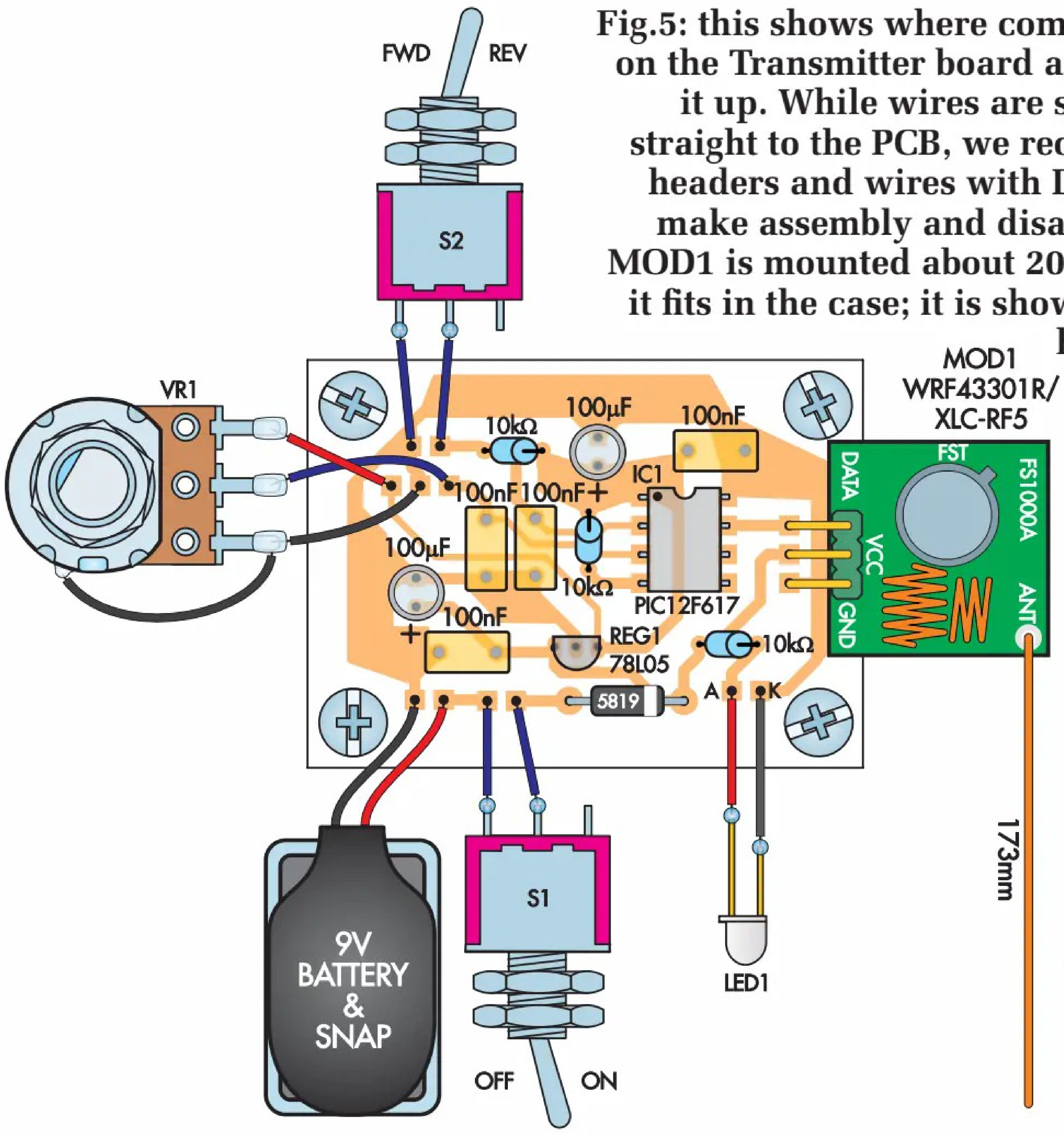


Fig.5: this shows where components mount on the Transmitter board and how to wire it up. While wires are shown soldered straight to the PCB, we recommend using headers and wires with DuPont plugs to make assembly and disassembly easier. MOD1 is mounted about 20° off vertical so it fits in the case; it is shown horizontally here for clarity.

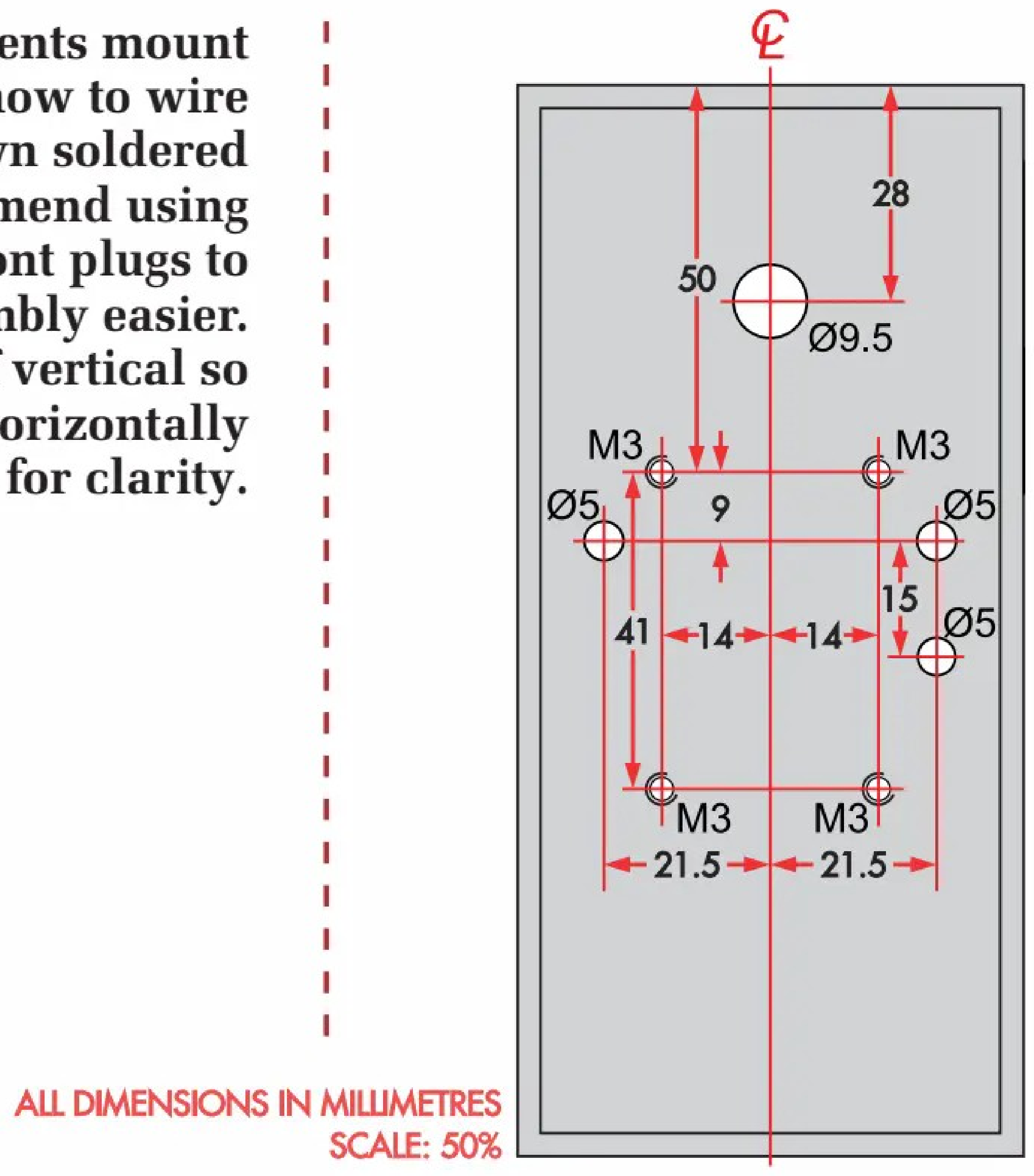


Fig.6: this view from the inside of the case front shows where to drill the holes. The large one is for the pot shaft, the 5mm holes are for the two switches and LED, while the M3-tapped holes are for mounting the board. If you would rather not tap them, drill them to 3mm and use extra machine screws (ideally countersunk) from the outside to fix the tapped spacers.

and an adaptor socket. For the former, you can use a PICkit or Snap programmer (or similar); for the latter, see our PIC Programming Adaptor (September 2024). Finally, check for any dry solder joints or solder bridges.

Case preparation

Drill and tap the Hammond 1593Y case as shown in Fig.6. That shows a view from the inside of the front part of the case. The large (9.5mm) hole is for the shaft of VR1, the three 5mm holes are for the two switches and LED, and the four M3-tapped holes are for mounting the PCB.

If you would rather not tap the holes, you can simply drill 3mm holes and use screws from both sides (which is accounted for in the parts list), but it will look worse and the extra screws will protrude outside the case unless you countersink them.

Now refer to Fig.5 and Photo 3 to see how everything goes together. Fit the LED, PCB, potentiometer, knob and toggle switches as shown. Split off the DuPont cables into sets, cut them in half, then solder them to the chassis-mounted parts and battery clip, using 1.5mm diameter heatshrink tubing to insulate the joints where necessary.

Solder a 173mm length of wire to the aerial pad on the transmitter module and insulate the other end. Then plug the DuPont connectors into the headers on the board using Fig.5 as a reference. Make sure everything goes to the right location, or it won't work properly.

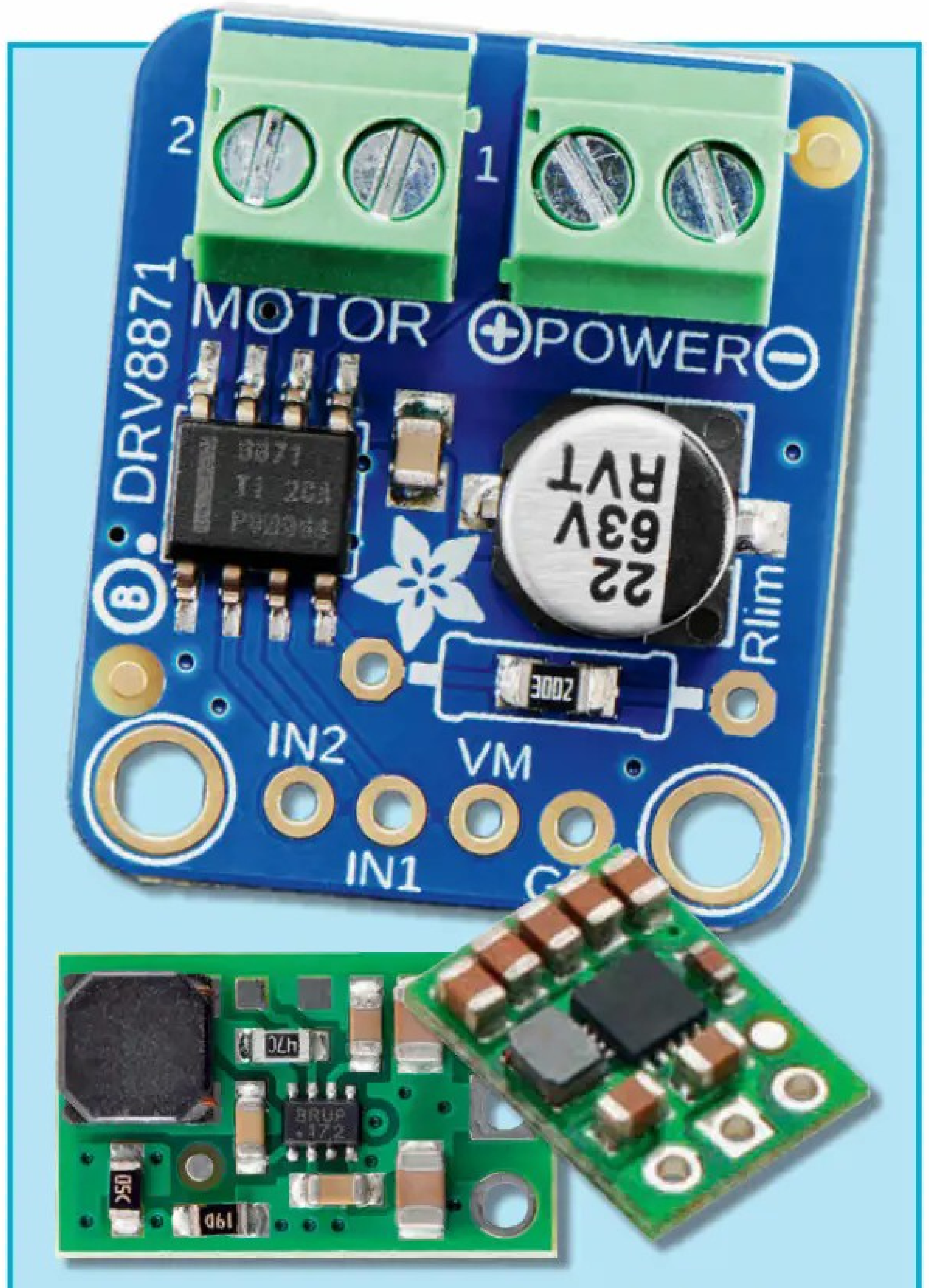


Photo 5: these are the Adafruit DRV8871 (top), Polulu U3V16F15 (lower left) and S7V7F5 (lower right) modules. We recommend you solder the right-angle headers so that they are parallel with the board (see Fig.2 and Photo 7).

Testing the Transmitter

With the microcontroller (IC1) out of its socket, check the orientation of the battery connector, 78L05 voltage regulator and the 433.9MHz transmitter module. Connect the 9V battery and switch it on. The LED on the front panel should glow. Connect a voltmeter with its red probe to pin 1 on the IC socket and the black lead to pin 8. The measured voltage should be very close to +5V DC.

If not, verify that the 5V regulator is the correct way round and there aren't any solder bridges shorting any tracks or pins. Assuming it's OK, switch off the power and insert the microcontroller. If you have an oscilloscope, connect it to pin 7 of the IC, with the Earth connector to 0V.

Switch on and you should should be able to capture a serial data waveform at 1200 baud similar to that in Scope 1.

If all is good, attach the back of the case using the supplied screws and you are ready to move on to the Receiver.

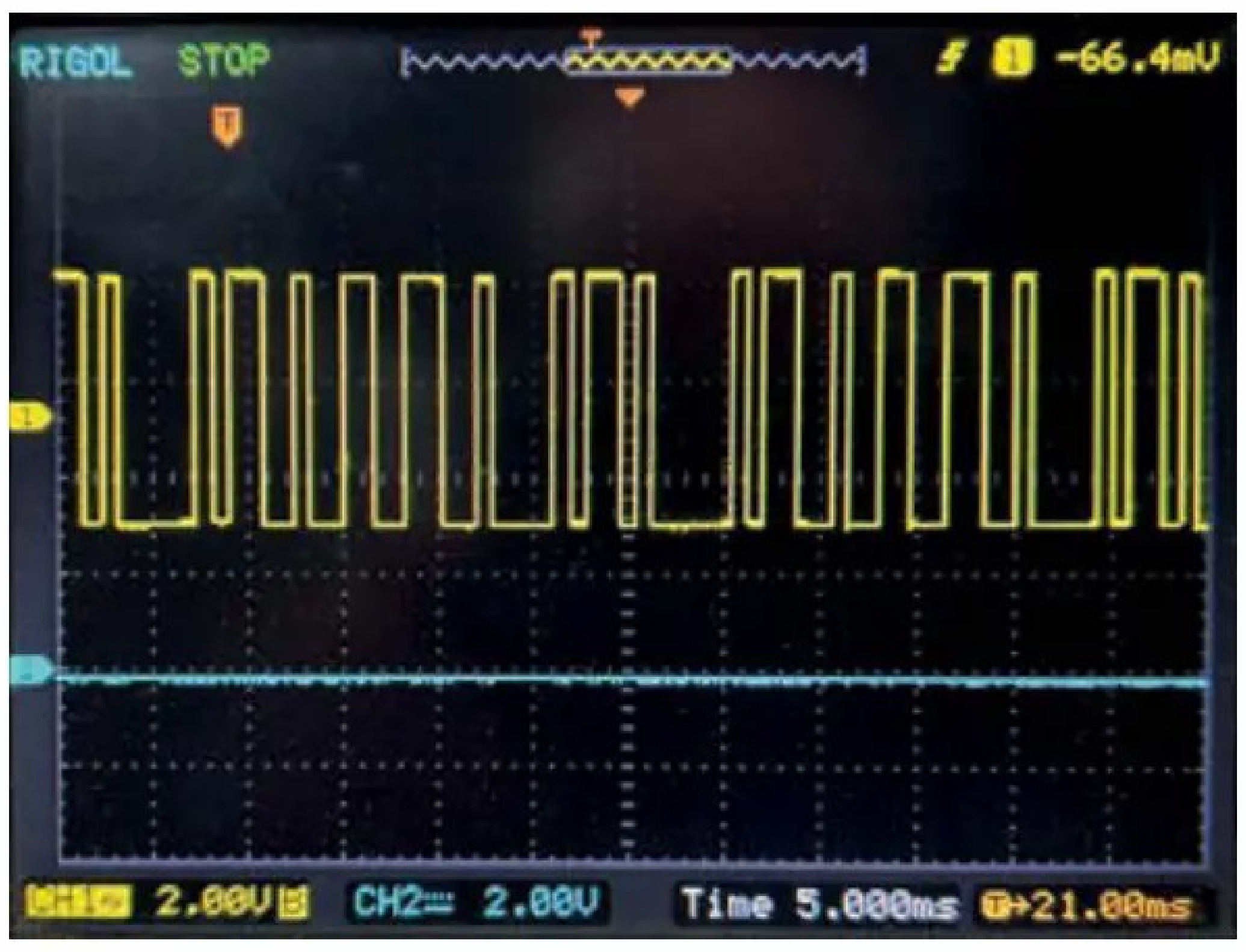
Receiver construction

First you must decide which version of the Receiver you want to build. The all through-hole version is larger at 74 × 23mm and uses a PCB coded 09110242, while the mixed SMD/TH version measures just 23 × 30mm with a PCB coded 09110243.

Both versions share many parts (all the modules are the same). The main difference is that the smaller version uses an SMD microcontroller and mostly SMD passives. The smallest parts are 2.0 × 1.2mm, so they are not terribly difficult to handle, and the IC has a fairly generous 1.27mm lead pitch.

The surface-mount PCB is the one I used to fit in my 85mm-long OO gauge carriage. You will need to use the SMD version if the TH board won't fit in yours; otherwise, the choice is yours.

The first task for both types of PCBs is to solder the supplied header pins to both of the Polulu DC/DC converter modules. Assemble them as shown in Fig.2, Fig.7 and Photo 6, making sure that the pins are parallel with the module PCBs. For the DVR8871 module, you have to add a four-pin right-angled header; again, make sure that the pins are parallel with the DVR8871 PCB.



Scope 1: this shows the serial data that's transmitted via a 433.9MHz wireless link with the switch in the forward position and the speed control at about halfway.

SMD PCB assembly

Since I etched mine myself, it is a single-sided design, although you can get the double-sided version from Silicon Chip, which avoids the need to fit the two wire links. The surface-mount components go on the copper side of the board, while the though-hole components and modules are inserted from the opposite side. The overlay diagram (Fig.7) shows both sides.

This is a good project if you are interested in improving your SMD soldering skills, since it has a few different types and sizes of components. I am 79 and can still manage these parts.

The SOIC-package PIC16F1455 will need to be programmed at some point. The easiest way is to purchase a pre-programmed PIC, although it is possible to program it in-circuit. See the panel for details if you wish to do that.

Use a flux pen or a syringe of flux paste to coat the PIC16F1455 IC's leads and its associated pads. Hold the PIC in place (eg, using tweezers) with the correct orientation and use your soldering iron to tack solder one lead in place, then check that it is positioned correctly. If so, solder the remaining leads.

Clean off the flux residue and inspect the leads under magnification to ensure that all the solder joints have formed correctly. If you are not sure about any of them, add more flux and apply heat (and possibly more solder) to reflow the joint. If you have bridged any pins, use more flux and some solder wick to remove the excess solder.

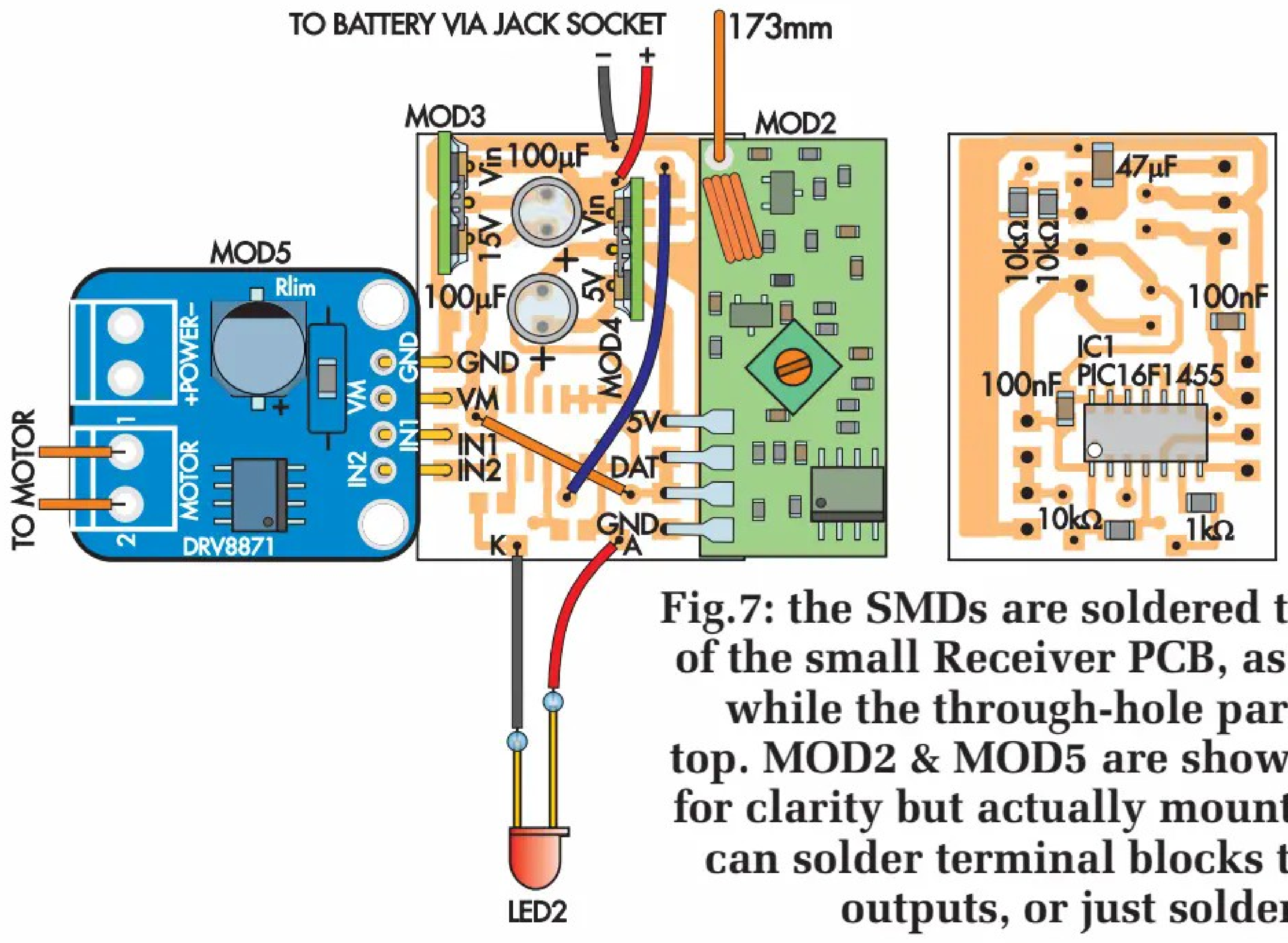


Fig.7: the SMDs are soldered to the underside of the small Receiver PCB, as shown at right, while the through-hole parts mount on the top. MOD2 & MOD5 are shown on their sides for clarity but actually mount vertically. You can solder terminal blocks to MOD5 for the outputs, or just solder wires directly.

Now use a similar procedure to fit the remaining SMDs. They are all the same size except the 47µF capacitor, which is a bit larger. The 1kΩ resistors will have a code like 102 or 1001 printed on top, while the code for 10kΩ is 103 or 1002. The capacitors will not be labelled.

Finally, using an ohmmeter on its lowest range, check each passive SMD component across its terminals to make sure you haven't accidentally created any short circuits.

Turn the board over and solder in the links (if you are using a single-sided board), the two electrolytic capacitors, and the four modules. Make sure all the components are the right way around. The four modules are mounted at right-angles to the main board, although some are shown horizontally in Fig.7 for clarity.

The final task is to attach the headers and connect the wires to the red LED and train motor. Disconnect the wires that connect the train wheels to the motor because we don't want the rails to act as aerials to radiate interference from the motor brushes.

For my 85mm carriage, the motor wires are 12cm long, the wires from the PCB to the connector are 7cm long, the wires from the jack plug to the PCB connector are 6cm long, the wires from the jack plug to the battery connector are 6cm long and the battery connector wire is 4cm long. All connections are insulated using heatshrink tubing.

Inside the train engine, the manufacturer should have fitted two inductors in series with the motor wires (typically around 30µH) and a 100nF capacitor across the motor terminals to suppress radiation from the motor brushes often on a small PCB.

It is important to have such a circuit, as without it, the radiated signal can be picked up by the receiver, causing potentially erratic operation. If it is missing, the train's manufacturer should be able to supply a new one.

The wires to the wheels should be disconnected from the two series inductors. The engine is powered from the carriage by a twisted pair of thin cable that connects from a two-pin male header to the two series inductors inside the engine.

Finally, connect a 173mm length of multi-stranded wire to the antenna terminal of the receiver module. All connections should be insulated using heatshrink tubing.

SMD version testing

Connect a voltmeter between the LED anode (red) wire and the 0V battery input, and a dual-trace oscilloscope to IN1, IN2 with the Earth connected to the 0V input. Connect a variable power supply to the 4.8V battery input, with the red wire going to the positive terminal and the black to the ground terminal. Slowly increase the voltage to about 5V; the meter should read 5V.

Switch on the Transmitter with the speed control set about halfway. The oscilloscope should show a 5V peak-to-peak 7kHz waveform with about a 50% duty cycle on either IN1 or IN2 (depending on the position of the forward/reverse switch). Increase the speed to maximum, and the display should change to a continuous +5V DC. On reducing it to minimum, you should see a 6% duty cycle square wave.

If IN1 shows the 7kHz waveform then IN2 should be at +5V, while if IN2 shows the 7kHz waveform, IN1 should be at +5V. Reduce the input voltage to less than 4V and you should see the red LED switch on.

If you don't have an oscilloscope, you can instead connect a DVM to either IN1 or IN2 (with the black probe to ground) and vary the speed potentiometer. The DVM should read the average voltage of the PWM signal, meaning it should increase smoothly as you advance the speed control clockwise. If it's stuck at 5V, switch the DVM probe to the other terminal (IN1 or IN2).

Through-hole version

If you have long OO gauge carriages or a train that will take the

Fig.8: the larger Receiver board uses all through-hole parts that mount on the top. You only need to fit the three wire links if you have a single-sided board. All modules mount vertically; MOD5's component side is towards the bottom of the PCB as shown, while MOD2 has the majority of its components near the top edge.

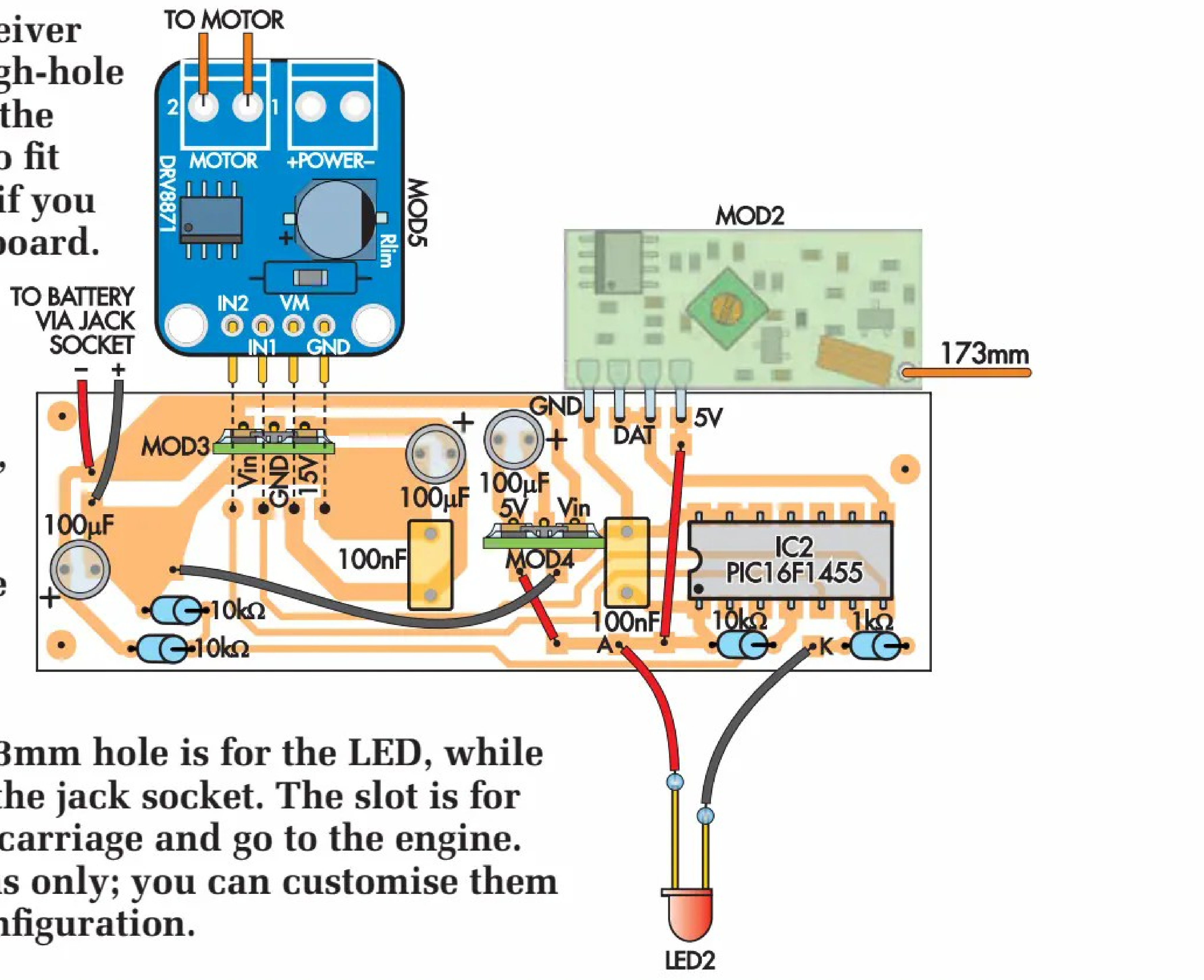


Fig.9 (far right): the 3mm hole is for the LED, while the 4mm hole is for the jack socket. The slot is for the wires to exit the carriage and go to the engine. These are suggestions only; you can customise them for your carriage configuration.

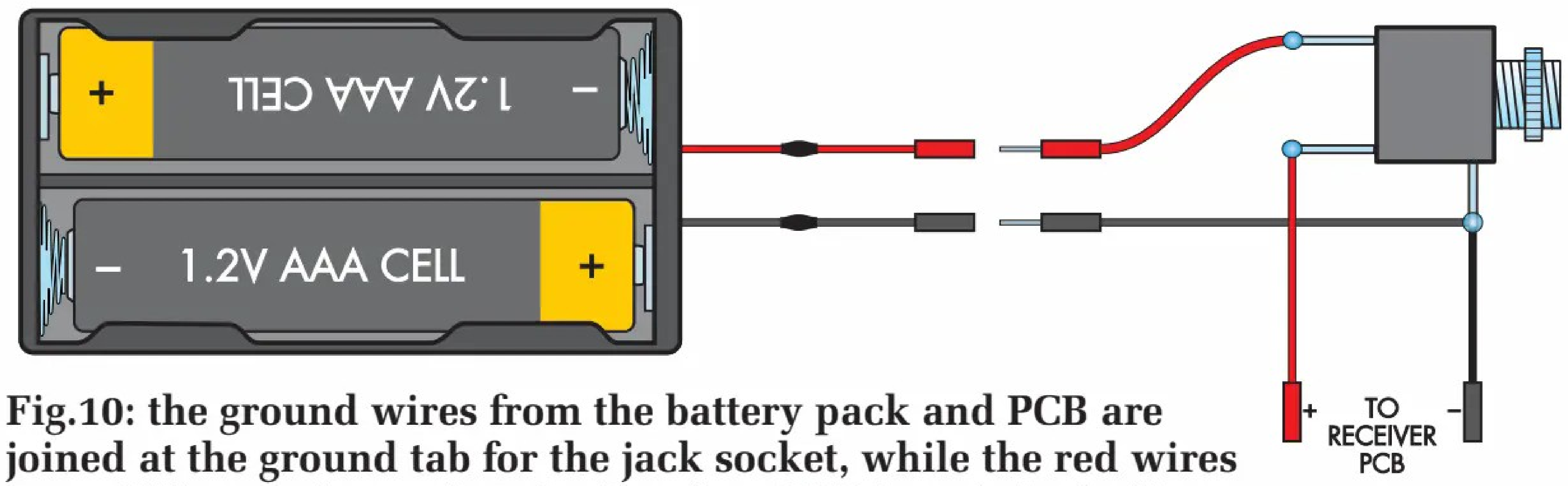


Fig.10: the ground wires from the battery pack and PCB are joined at the ground tab for the jack socket, while the red wires go to different pins so that the Receiver PCB is switched off when the jack plug is inserted (for charging, or just to cut the power).

board and batteries, you might find building this one a bit easier. Since I etched mine myself, it is a single-sided design, although you can get the double-sided version from Silicon Chip, which avoids the need to fit the wire links. Refer to the PCB overlay diagram, Fig.8.

Solder in the links (if you are using a single sided board), the three electrolytic capacitors, the 14-pin IC socket and DC/DC converter modules, making sure they are orientated correctly. Then add the headers, MKT/ceramic capacitors and resistors.

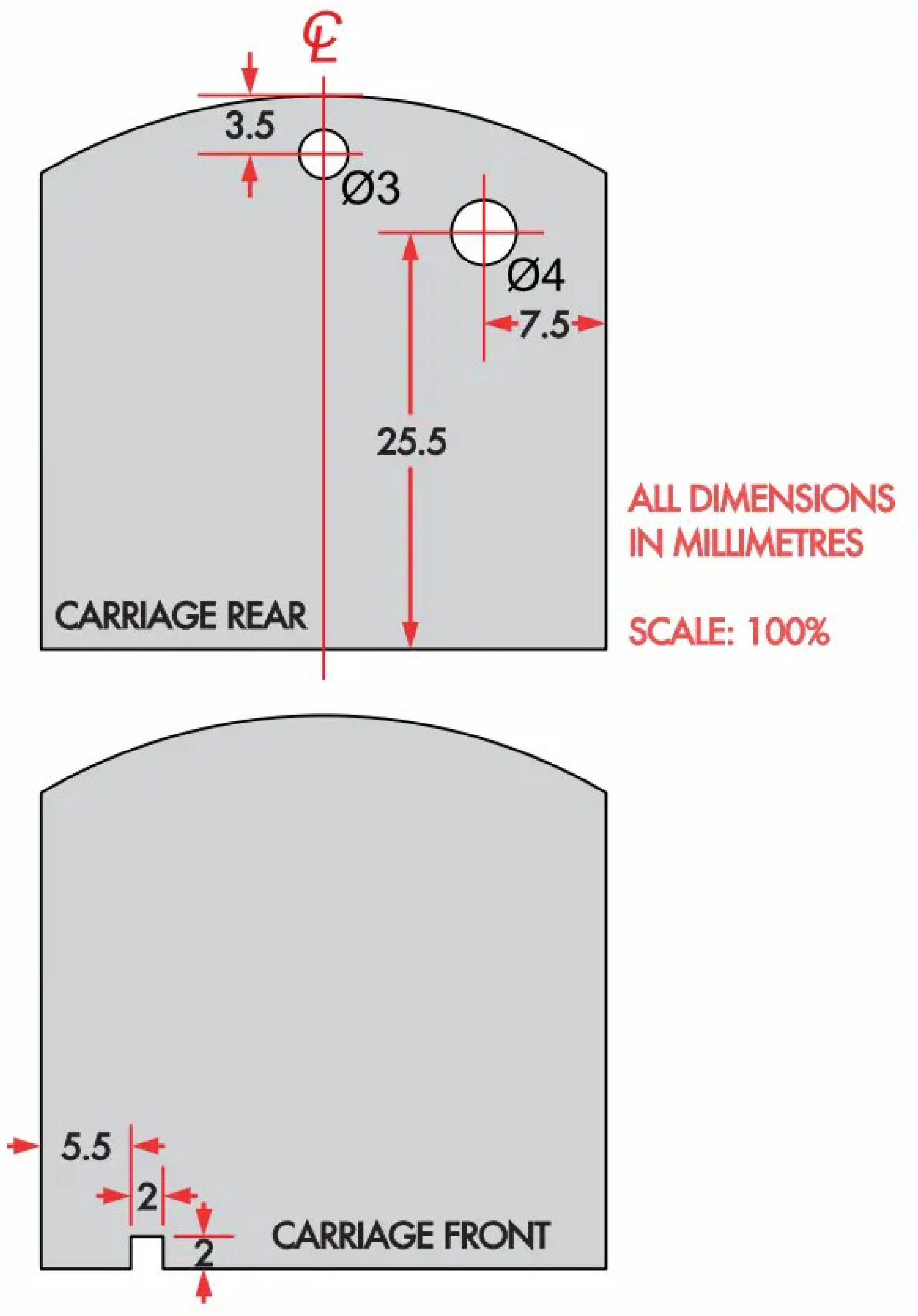
Wire up the red LED and train motor as shown. The length of the wires will depend on the size of the carriage you are using. All connections should be insulated by using heatshrink tubing.

Through-hole version testing

Check that the components are the correct way round and there are no solder bridges on the PCB.

Connect the battery red wire to the positive terminal of a 5V power supply and the black wire to the 0V terminal, switch it on and use a DVM to measure the voltage between pin 1 and pin 14 of the IC socket. It should be very close to 5V. Also check the 15V supply by measure the voltage between the V_{OUT} and 0V terminals of the U3V16F15 module. The result should be very close to 15V.

If all is well, fit the DVR8871 H-bridge module, 433.9MHz receiver and insert the PIC16F1455 chip into its socket, making sure they are all orientated correctly. Connect a 173mm length of multi-stranded wire to the antenna terminal of the receiver.



The rest of the testing is the same as that for the surface-mount version of the PCB, so refer to that section above. The wiring lengths are different for this version, as is the position of the red LED and jack plug socket. These will depend on your train's dimensions.

Mounting the Receiver

The 3mm LED and 2.5mm jack socket need to be mounted on the carriage, along with an access groove for the cable connecting to the engine. Fig.9 shows the suggested carriage cover modifications to achieve this. Fit the jack plug socket into the 4mm hole so that pin 1 is as close as possible to the side of the carriage cover.

Once they are mounted, wire up the jack socket and battery as shown in Fig.10. Insulate any exposed connections with 1.5mm diameter heatshrink tubing.

Next, load the battery holder with fully charged cells and connect the battery to the jack socket. Connect the black lead of a DVM to the negative wire that will go to the Receiver PCB in the engine, and the red lead to the positive wire. You should get a reading close to 4.8V (the charged battery voltage).

Now plug a jack plug into the socket and check again; the voltmeter should read 0V. Next, measure the voltage across the jack plug terminals and it should be once again be close to 4.8V.

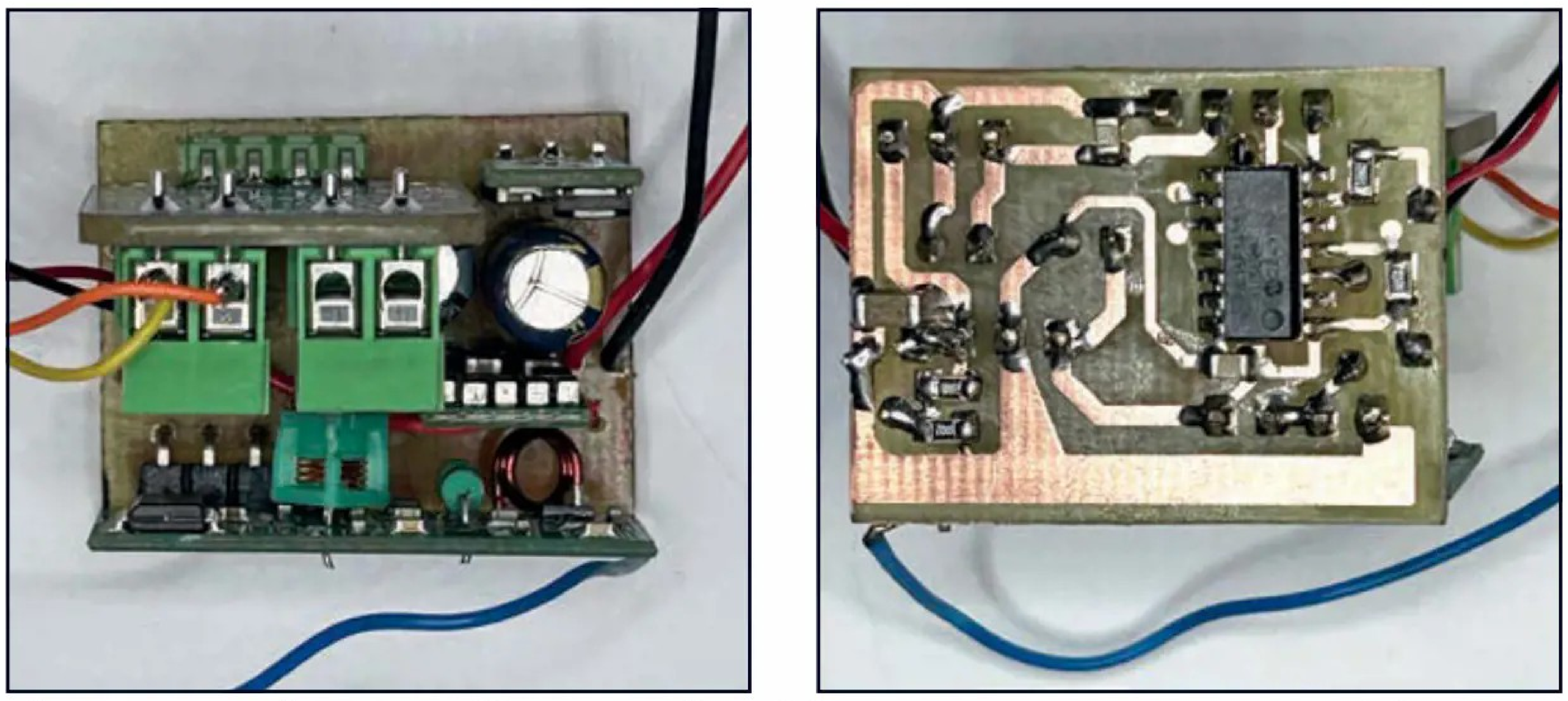
Insert the red LED into the 3mm hole. Fit the PCB and battery holder as shown in Photo 9. Connect the power wires to the Receiver PCB, tucking them and the excess wire down the side of the battery holder.

Coil the antenna cable and tuck it down between the PCB and the carriage end that holds the jack socket. Leave the jack plug in, as this stops power from the batteries flowing into the Receiver.

Cover the wheel assembly with a strip of insulating tape where the bottom of the PCB may contact it. You can then fit the wheel assembly to the carriage cover.

Final testing

Switch on the Transmitter and set the speed control to its minimum position. With the engine laying on its back, connect it to the carriage. Switch on the Receiver by removing the jack plug. Rotate the speed control on the Transmitter and the engine wheels should start to move, gaining speed as the control is rotated further until maximum speed is reached.



Photos 6 & 7: the top and bottom sides of the prototype SMD version of the Battery-Powered Model Train Receiver PCB.

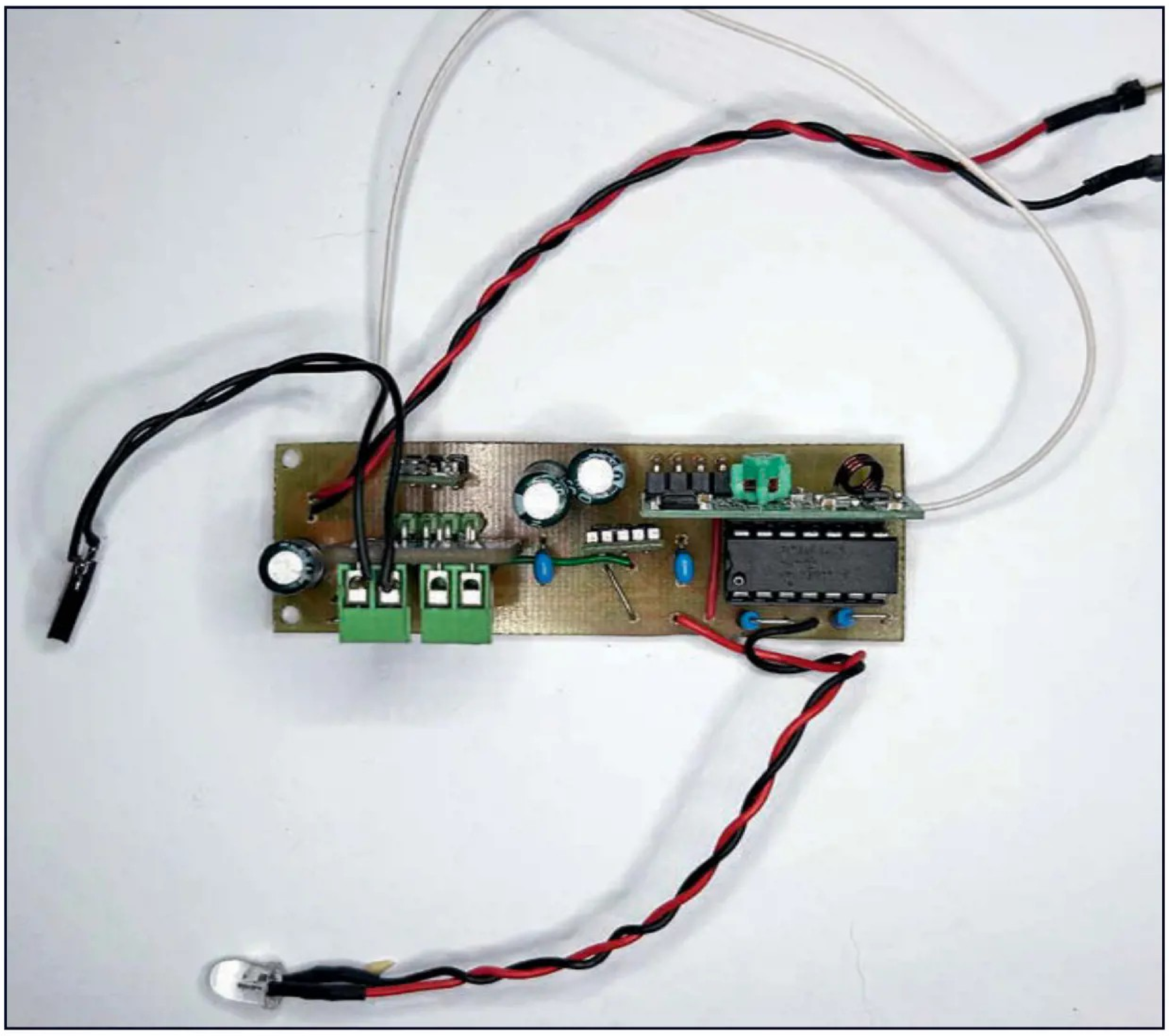


Photo 8: the through-hole version of the Receiver PCB is much larger than the SMD version (about twice as wide), but it is easier to assemble due to using through-hole components.

Photo 9: the SMD Receiver PCB and four AAA cells just fit into a OO-gauge train carriage.

Turn the control back down and the speed should decrease to zero just before minimum rotation. Repeat with the forward/reverse switch in the other position. Switch off the Transmitter and insert the jack plug to switch off the Receiver.

Running the train

Place the engine and assembled carriage onto the tracks and connect the motor lead and socket. Switch on the Transmitter and turn the speed control to minimum and the forward reverse switch to forward. Remove the jack plug from the carriage (power on). Rotate the potentiometer clockwise and the train should move forward; its speed should increase with the advancement of the control.

If it goes in reverse, unplug the motor leads from the train and reverse the connections. It should now run forwards.

Repeat the test with the switch in the reverse position. With the train running, switch off the Transmitter; the train should continue running at the same speed. Switch the Transmitter on again, rotate the potentiometer fully anti-clockwise and the train should stop.

Insert the jack plug to switch the Receiver off. If the red LED is lit, plug in the Charger until the batteries are charged.

Always stop the train before operating the forward/reverse switch; failure to do so may destroy the motor. Always switch the Transmitter on before switching the train on, and always switch off the train off before the Transmitter.

This avoids the train running by itself if in the unlikely event of an interfering signal that's interpreted as valid by the Receiver.

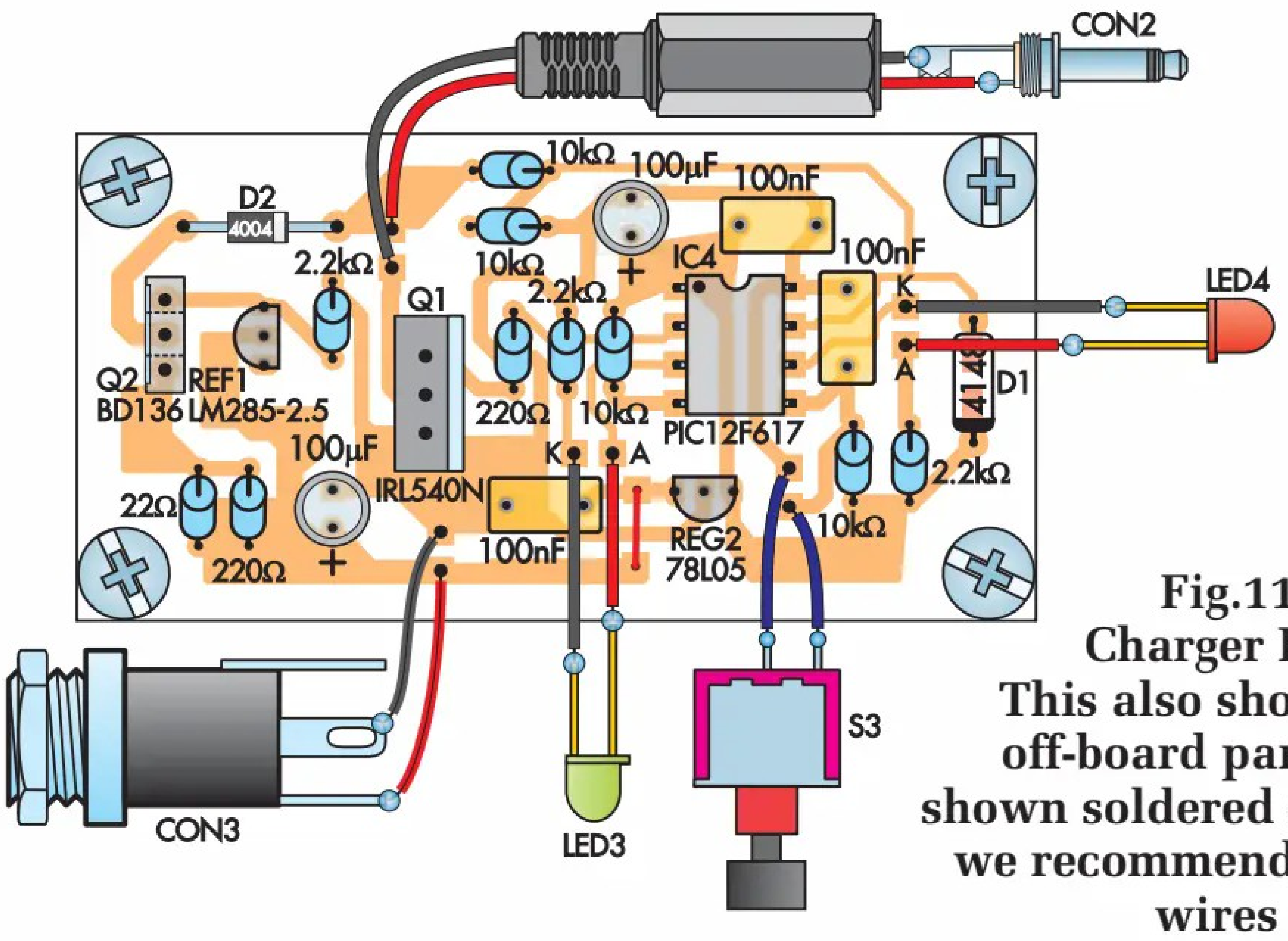


Fig.11: fit the parts to the Charger PCB as shown here. This also shows how to wire the off-board parts. While wires are shown soldered straight to the PCB, we recommend using headers and wires with DuPont plugs.

Charger construction

The Charger is built on a single- or double-sided PCB coded 09110244 that measures 63 × 32mm. Its overlay diagram is shown in Fig.11. Once again, headers are not shown in the wiring but it's easiest to use headers and plugs. Start by fitting the headers, IC socket, wire link (if needed) and the capacitors. Take care to orientate the socket and electrolytic capacitors correctly.

...continued on page 76

Practical Electronics PCB SERVICE

PROJECT	CODE	PRICE
DECEMBER 2025		
Variable Speed Drive (black)	11111241	£11.99
Digital Capacitance Meter	04111241	£5.99
Battery-Powered Model Train transmitter	09110241	£3.99
Battery-Powered Model Train through-hole receiver	09110242	£3.99
Battery-Powered Model Train SMD receiver	09110243	£3.99
Battery-Powered Model Train battery charger	09110244	£3.99
Transmanium Fuzz Box jack PCB	AO-DEC-25	£3.99
NOVEMBER 2025		
Capacitor Discharger	9047-01	£4.99
5MHz, 40A Current Probe	9049-01	£5.99
Transmanium Fuzz Box main PCB	AO-NOV-25	£4.99
OCTOBER 2025		
Seaside Sound Simulator PCB (blue)	01111241	£7.99
Fender Jazz Bass pickguard	23109241	£5.99
J&D Luthiers T-Style pickguard	23109242	£5.99
Music Man Stingray pickguard	23109243	£5.99
Fender Telecaster pickguard	23109244	£4.99
3D Printer Filament Dryer control PCB	28110241	£4.99
SEPTEMBER 2025		
Automatic LQ Meter PCB	CSE240203A	£5.99
Automatic LQ Meter front panel (black)	CSE240204A	£5.99
Compact Hi-Fi Headphone Amplifier PCB (blue)	01103241	£4.99
Discrete Ideal Bridge Rectifier TH PCB	18108241	£2.99
Discrete Ideal Bridge Rectifier SMD PCB	18108242	£2.99
Dual Rail Supply Protector PCB	18109241	£3.99
Audio Out Cascode Amplifier PCB	CASCODEA	£4.99
Audio Out Cascode Amplifier Power Supply PCB	CASCODEP	£4.99
AUGUST 2025		
Micromite Explore-40 PCB	07106241	£2.99
Micromite Explore-40 Audio Breakout PCB	07101222	£2.99
8-Channel Learning IR Remote Receiver PCB	15108241	£5.99
FlexiDice PCB set	08107241-2	£4.99
JULY 2025		
Compact OLED Clock & Timer PCB	19101231	£4.99
180-230V DC Motor Speed Controller PCB	11104241	£9.99
Mains Power-Up Sequencer PCB (same as Feb '25)	10108231	£11.99
JUNE 2025		
Styloclone (case version)	23106241	£6.99
Styloclone (standalone version)	23106242	£7.99
Dual Mini LED Dice (TH LEDs, white solder mask)	08103241W	£2.99
Dual Mini LED Dice (TH LEDs, black solder mask)	08103241B	£2.99
Dual Mini LED Dice (SMD LEDs, white solder mask)	08103242W	£2.99
Dual Mini LED Dice (SMD LEDs, black solder mask)	08103242B	£2.99
Dual Hybrid Supply regulator board	18107211	£5.99
Dual Hybrid Supply front panel control board	18107212	£2.99
Dual Hybrid Supply CPU board	01106193	£3.99
Dual Hybrid Supply LCD adaptor	01106196	£2.99
MAY 2025		
Skill Tester 9000	08101241	£9.99
USB-C Serial Adaptor (black solder mask)	24106241	£2.99
DC Supply Protector (SMD adjustable version)	08106241	£3.99
DC Supply Protector (TH adjustable version)	08106242	£3.99
DC Supply Protector (TH fixed version)	08106243	£3.99
APRIL 2025		
ESR Test Tweezers PCB set	04105241+	£5.99
Fan Speed Controller Mk2	10104241	£4.99
Calibrated Microphone (SMD version)	01108231	£3.99
Calibrated Microphone (TH version)	01108232	£3.99
MARCH 2025		
Pico Gamer	08104241	£9.99
Compact Frequency Divider	04112231	£4.99
FEBRUARY 2025		
Laser Communicator transmitter	16102241	£3.99
Laser Communicator receiver	16102242	£3.99
Model Railway Points Controller transmitter	09101241	£3.99
Model Railway Points Controller receiver	09101242	£2.99
Mains Power-Up Sequencer	10108231	£11.99
JANUARY 2025		
USB to PS/2 Converter for VGA PicoMite	07111231	£3.99
ps2x2pico USB to PS/2 Converter	07111232	£3.99
Compact Microphone Preamplifier (case version)	01110231	£5.99
Compact Microphone Preamplifier (integrated version)	01110232	£5.99
DECEMBER 2024		
Raspberry Pi Clock Radio PCB pair	19101241/2	£12.99
VGA PicoMite (blue)	07107221	£5.99
Secure Remote Switch receiver	10109231	£4.99
Secure Remote Switch transmitter (module version)	10109232	£2.99
Secure Remote Switch transmitter (discrete version)	10109233	£2.99
NOVEMBER 2024		
Ideal Diode Bridge Rectifiers (28mm square)	18101241	£3.99
Ideal Diode Bridge Rectifiers (22mm square)	18101242	£3.99
Ideal Diode Bridge Rectifiers (4-pin SIL)	18101243	£3.99
Ideal Diode Bridge Rectifiers (tiny SOT-23)	18101244	£3.99
Ideal Diode Bridge Rectifiers (standalone SMD)	18101245	£4.99
Ideal Diode Bridge Rectifiers (standalone TH; 2oz copper)	18101246	£6.99
Multi-channel Volume Control (volume PCB)	01111221	£4.99
Multi-channel Volume Control (control PCB)	01111222	£4.99
Multi-channel Volume Control (OLED PCB, black)	01111223	£4.99
OCTOBER 2024		
TQFP-32 Programming Adaptor	24108231	£4.99
TQFP-44 Programming Adaptor	24108232	£4.99
TQFP-48 Programming Adaptor	24108233	£4.99
TQFP-64 Programming Adaptor	24108234	£4.99
2m VHF FM Test Signal Generator	06107231	£4.99
SEPTEMBER 2024		
PIC Programming Adaptor	24105231	£4.99
Salad Bowl Speaker crossover	01109231	£7.99
AUGUST 2024		
Arduino ESR Meter combined L/C-ESR	04106182	£9.99
Arduino ESR Meter separate ESR	04106183	£7.99
Watering System Controller	15110231	£11.99
JULY 2024		
Dynamic NFC Tag (22 × 31mm)	06101231	£5.95
Dynamic NFC Tag (86 × 54mm)	06101232	£7.95
Reciprocal Frequency Counter	CSE230101C	£8.95
Reciprocal Frequency Counter front panel (black mask)	CSE230102	£7.95
Pi Pico-based Thermal Camera	04105231	£6.95
JUNE 2024		
Loudspeaker Test Jig	04106231	£12.95
MAY 2024		
GPS-Disciplined Oscillator	04103231	£12.95
Dual RF Amplifier	CSE220602A	£5.95
Songbird	08103231	£7.95

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PCBs for most recent *PE/EPE* constructional projects are available. From the July 2013 issue onwards, PCBs with eight-digit codes have silk screen overlays and, where applicable, are double-sided, have plated-through holes, and solder mask. They are similar to photos in the project articles. Earlier PCBs are likely to be more basic and may not include silk screen overlay, be single-sided, lack plated-through holes and solder mask.

Always check price and availability in the latest issue or online. A large number of older boards are listed for ordering on our website.

In most cases we do not supply kits or components for our projects. For older projects it is important to check the availability of all components before purchasing PCBs.

Magazine back issues are available – see https://www.electronpublishing.com/product-category/pe-back-issues-print-and-digital/

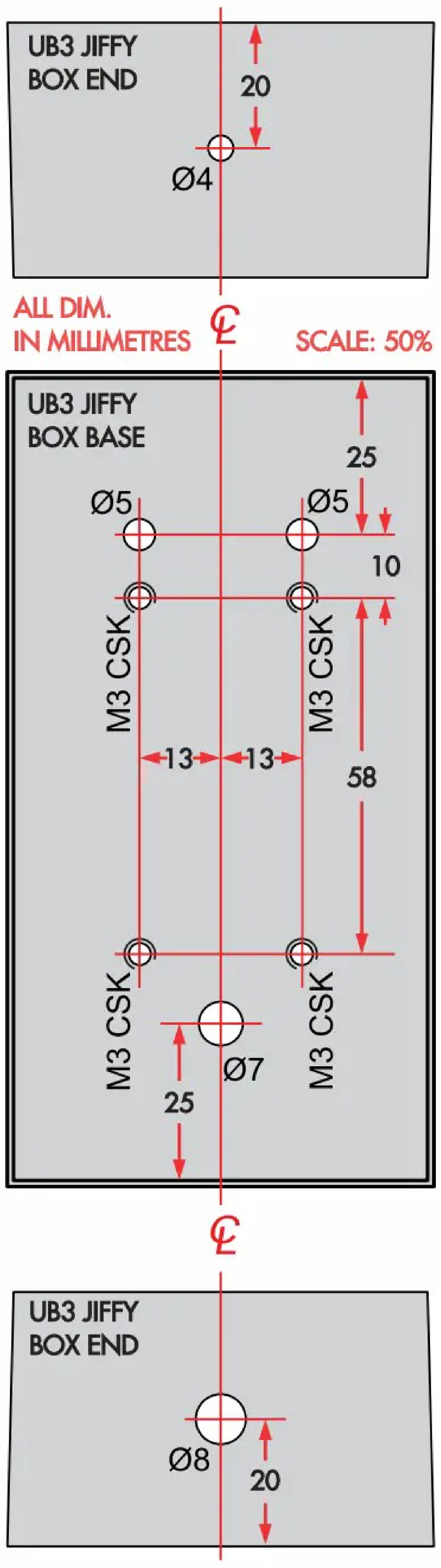

Fig.12: the Jiffy box needs holes at each end for the power input and charging output, plus four countersunk holes for mounting the PCB, plus three more for the pushbutton and two LEDs.

Parts List – Battery-powered Model Train

1 500mm length of 1.5mm diameter black or clear heatshrink tubing
various lengths & colours of light-duty hookup wire
(wire for the power to the engine can be from old USB and mouse cables)

Charger

1 single- or double-sided PCB coded 09110244, 63 × 32mm
1 UB3 Jiffy box
1 9V DC 150mA+ plugpack
1 2.5mm mono jack plug (CON2) [Jaycar PP0100]
1 chassis-mount DC socket to suit plugpack (CON3)
1 chassis-mount SPST miniature momentary pushbutton (S3)
1 8-pin DIL IC socket
5 2-way pin headers, 2.54mm pitch
6 female-female DuPont jumper wires, ideally joined in a ribbon
4 M3 × 8mm countersunk head machine screws
8 M3 hex nuts
1 500mm length of single-core screened microphone cable
1 PIC12F617-I/P 8-bit microcontroller programmed with 0911024C.HEX, DIP-8 (IC4)
1 LM285-2.5 voltage reference diode, TO-92 (REF1)
1 78L05 5V 100mA linear regulator, TO-92 (REG2)
1 IRL540N 100V 36A Mosfet, TO-220 (Q1)
1 BD136/138/140 45/60/80V 1.5A PNP transistor, TO-126 (Q2)
1 5mm green LED (LED3)
1 5mm red LED (LED4)
1 1N4004 400V 1A diode (D2)
1 1N4148 75V 200mA diode (D3)
2 100µF 16V low-ESR radial electrolytic capacitors
3 100nF 50V ceramic, multi-layer ceramic or MKT capacitors
4 10kΩ ¼W 1% axial resistors
3 2.2kΩ ¼W 1% axial resistors
2 220Ω ¼W 1% axial resistors
1 39Ω 1W 1% axial resistor (for testing)
1 22Ω ¼W 1% axial resistor

Transmitter

1 single- or double-sided PCB coded 09110241, 49 × 36mm
1 Hammond 1593Y plastic case **[DigiKey, Mouser, RS]**
1 3-pin 433.9MHz transmitter module, WRF43301R or XLC-RF5 (MOD1) **[AliExpress, eBay]**
1 9V battery snap
1 9V battery
1 8-pin DIL IC socket
1 3-way pin header, 2.54mm pitch
4 2-way pin headers, 2.54mm pitch
7 female-female DuPont jumper wires, ideally joined in a ribbon

Now add the resistors, which are mounted vertically, the BD136 transistor, IRL540N Mosfet, LM285-2.5V voltage reference diode, 78L05 voltage regulator, plus the 1N4148 and 1N4004 diodes. Make sure all the semiconductors are correctly orientated and in the right places. Don't fit the PIC microcontroller yet.

If you purchased the micro from the Silicon Chip Shop, it will already have the firmware loaded. If you wish to do this yourself, the files can be downloaded from siliconchip.au/Shop/6/508 and we had some comments earlier about ways to program the chip.

Once the PCB is fully assembled, check for any dry solder joints or solder bridges. It mounts in a UB3 Jiffy box that has to be drilled for the LEDs, pushbutton, PCB mounting screws and power input socket. The drilling details are shown in Fig.12.

Once the box has been drilled, attach the red and green LEDs, start pushbutton and the barrel socket as shown in the photos.

The PCB is held in place by four 8mm-long countersunk head M3 machine screws and eight M3 hex nuts. The four extra nuts are used to space the PCB off the case.

Use DuPont wires to make the connections between the PCB and the offboard components, as shown in Fig.11. Insulate all exposed connectors and the wire connections to the LEDs with 1.5mm diameter heatshrink tubing.

Finish the Charger off by preparing the box, as shown in Fig.12, then mounting the PCB and all the chassis-mounting parts to it.

Testing the Charger

Make sure that the microcontroller is not in its socket; at the same time,

2 SPDT subminiature toggle switches (S1, S2)
1 10kΩ 16mm linear potentiometer with large knob (VR1)
8 M3 × 6mm panhead machine screws
4 M3 × 6mm tapped hex spacers
1 PIC12F617-I/P 8-bit micro programmed with 0911024T.HEX, DIP-8 (IC1)
1 78L05 5V 100mA linear regulator, TO-92 (REG1)
1 high-intensity 5mm LED, white recommended (LED1)
1 1N5819 40V 1A schottky diode (D1)
2 100µF 16V low-ESR radial electrolytic capacitors
4 100nF 50V ceramic, multi-layer ceramic or MKT capacitors
3 10kΩ ¼W 1% axial resistors

Receiver (common to both versions)

1 4-pin 433.9MHz receiver module, WRF43301R or XLC-RF5 (MOD2)
  [AliExpress, eBay]
1 Polulu U3V16F15 15V output step-up DC/DC converter (MOD3)
  [uk.robotshop.com/products/pololu-15v-step-up-voltage-regulator-u3v16f15]
1 Polulu S7V7F5 5V output step-up/down DC/DC converter (MOD4)
  [uk.robotshop.com/products/step-up-step-down-voltage-regulator-s7v7f5]
1 Adafruit DRV8871 motor driver module (MOD5)
4 1.2V 900mAh NiMH AAA cells
1 2×2 AAA battery holder with flying leads
1 2.5mm mono switched chassis-mounting jack socket (CON1)
2 4-way right-angle pin header, 2.54mm pitch (for MOD2 & MOD5)
2 female-female DuPont jumper wires, ideally joined together
1 red 3mm LED (LED2)

Receiver (TH version only)

1 single- or double-sided PCB coded 09110242, 74 × 23mm
1 PIC16F1455-I/P micro programmed with 0911024R.HEX, DIP-14 (IC2)
1 14-pin DIL IC socket
3 100µF 16V low-ESR radial electrolytic capacitors
2 100nF 50V ceramic, multi-layer ceramic or MKT capacitors
3 10kΩ ¼W 1% axial resistors
1 1kΩ ¼W 1% axial resistor

Receiver (SMD version only)

1 single- or double-sided PCB coded 09110243, 23 × 30mm
1 PIC16F1455-I/SL micro programmed with 0911024R.HEX, SOIC-14 (IC2)
1 100µF 16V low-ESR radial electrolytic capacitor
1 100µF 6.3V radial electrolytic capacitor
1 47µF 16V X5R M3216/1206 SMD ceramic capacitor
2 100nF 50V X7R M2012/0805 SMD ceramic capacitors
3 10kΩ ⅛W 1% M2012/0805 SMD resistors
1 1kΩ ¼W 1% M2012/0805 SMD resistor

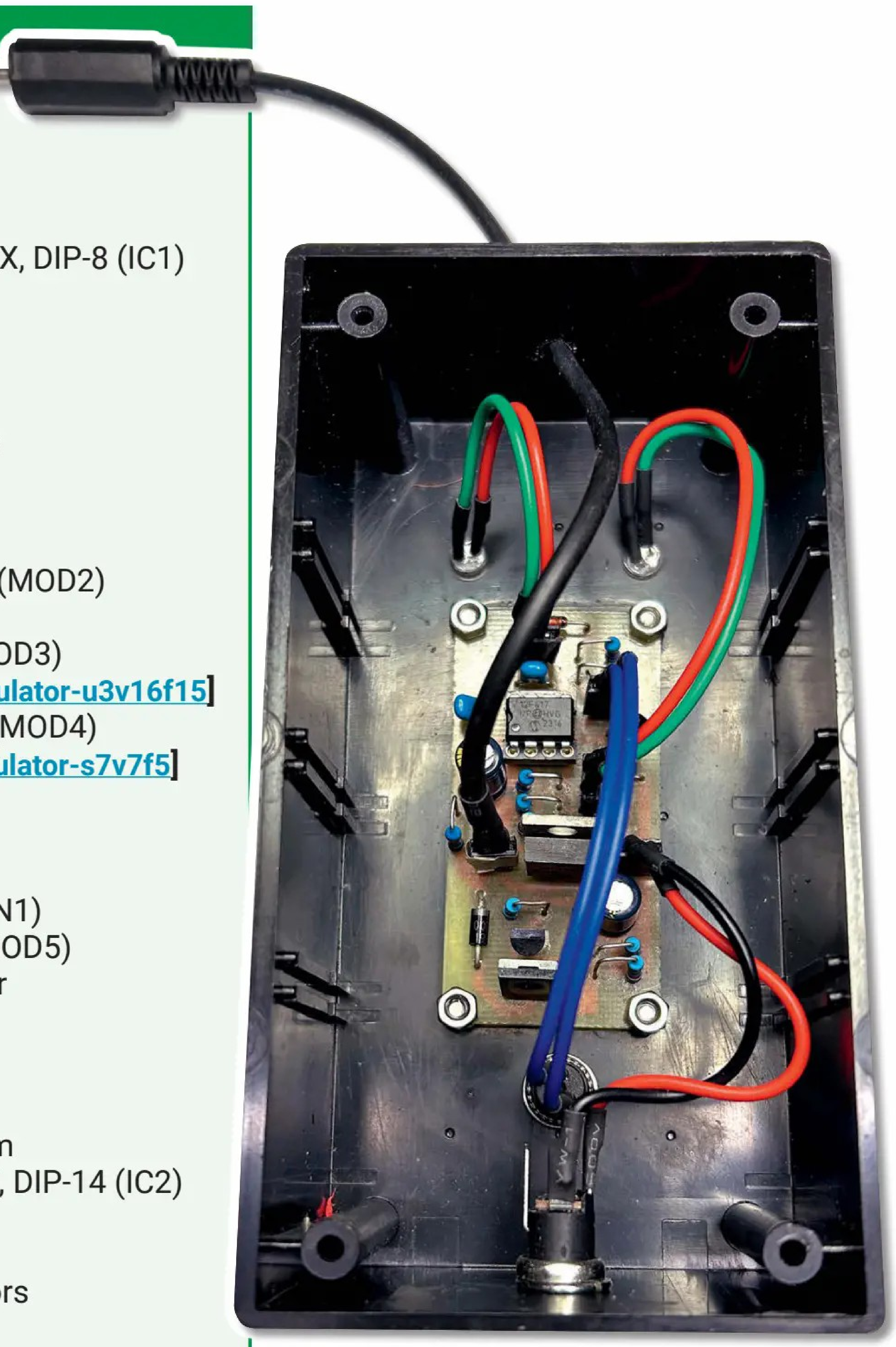

Photo 10: the Charger board easily fits inside a UB3 Jiffy box (or a smaller case) as shown here and in Photo 3.

check the orientation of all the semiconductors and electrolytic capacitors. Connect the power supply and switch it on. Take a voltmeter and connect the red lead connected to pin 1 of the empty IC socket, and the black lead to pin 8. You should measure very close to +5V DC.

If not, check that the 5V regulator is the correct way round and there aren't any solder bridges shorting the tracks.

Assuming it's OK, switch off the power, insert the microcontroller and connect a 39Ω 1W resistor between the battery terminals (eg, using clip leads). Apply power again and the green LED should flash.

Press the Start button; the green LED should extinguish and the red LED should flash, indicating 'charging'. There should be about 3.5V across the 39Ω resistor, indicating 90mA of current flow.

To simulate a fully charged battery, disconnect the 39Ω resistor. The green LED should then flash, and the red LED will extinguish. If you want to check that the timer is working, reconnect the 39Ω resistor, press the Start button again and wait for 16 hours. The red LED should extinguish and the green LED will flash.

Using the Charger

When the battery voltage in the carriage falls below 4V, the 3mm LED in the rear of the carriage glows, alerting you that the battery needs charging.

Connect the Charger to the carriage via the 2.5mm jack plug. Switch on the Charger and press the Start button to begin charging. The Charger will revert to standby mode (with the green LED flashing) when the battery is fully charged.

Double-sided | plated-through holes | solder mask

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Advanced SMD Test Tweezers legs (pair)	04106212	£7.95
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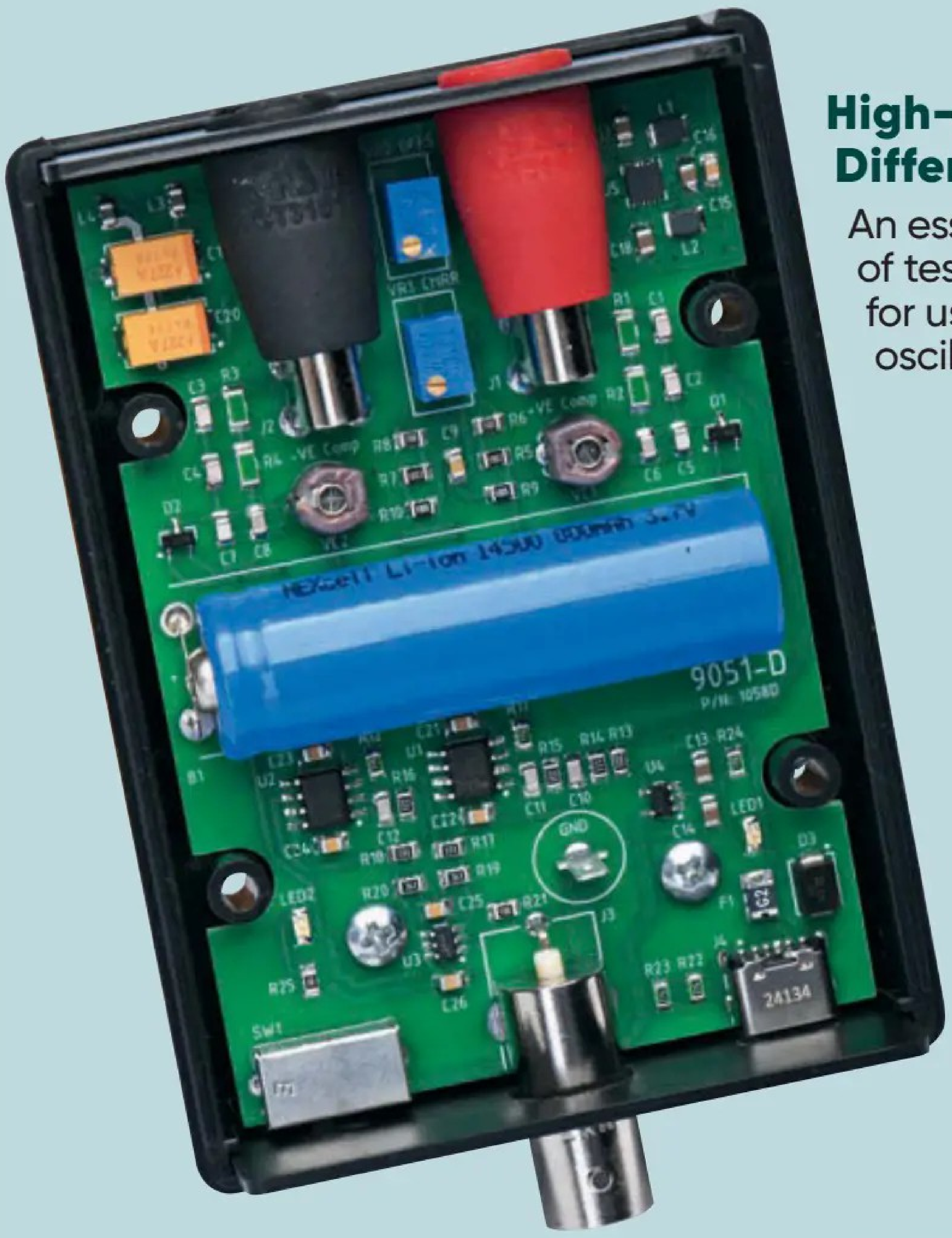

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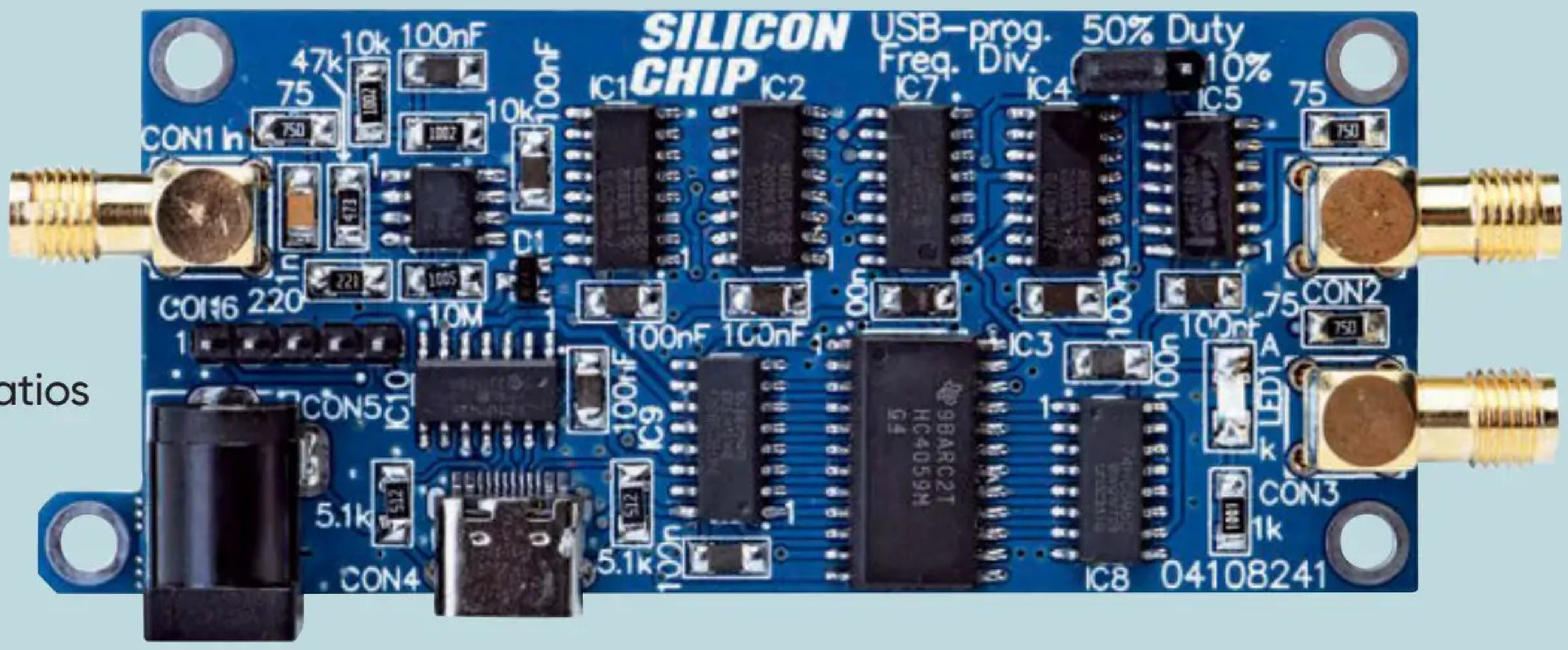

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